A SCREEN-GRID COUPLED

DIRECT-CURRENT AMPLIFIER

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e_g	Incremental grid voltage
e_o	Output voltage
e_o	Signal component of output voltage
e_p	Signal component of plate voltage
e_s	Signal input voltage
e_{sg}	Signal component of screen voltage
E_B	D-C plate supply voltage
E_C	D-C grid supply voltage
E_G	D-C grid bias voltage
g_m	Dynamic transconductance
g_{m2}	Dynamic screen-plate transconductance
i	An incremental current
i_p	Signal component of plate current
i_{sg}	Signal component of screen current
I	A direct current
I_p	Zero signal plate current
r_p	Dynamic plate resistance
r_{sg}	Dynamic screen resistance
R_b	Static plate resistance
v	Cathode drift voltage
μ	Dynamic amplification factor
μ_i	Dynamic screen-plate current amplification

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CHAPTER I

INTRODUCTION

I. DIRECT COUPLING

Direct-current amplifiers are required in circuits where the signal to be amplified is of zero or very low frequency. In such an amplifier, successive stages must be coupled directly or through resistance networks. In particular, blocking condensers cannot be used because they would block the desired signals as well as other direct potentials. In the literature and in the following discussion the terms direct-current amplifier and direct-coupled amplifier are used interchangeably.

II. USES FOR D-C AMPLIFIERS

In addition to the more common uses of amplifiers (such as increasing the sensitivity of meters or operating relays in automatic control circuits), d-c amplifiers may be used in circuits where the conventional amplifier would be of little or no use. D-C amplifiers can be operated directly from the output of a photoelectric cell. They permit the continuous recording of sound level, light intensity, and other phenomena easily converted into weak direct currents. Interest in the electrical properties of living organisms has brought about the development of very sensitive and stable amplifiers to record the minute potentials encountered. Amplifiers have been built with a sensitivity of a few microvolts and a voltage amplification of more than one million.[1]

III. OTHER CLASSES OF D-C AMPLIFIERS

There are, besides the class of amplifier to be considered here, two other classes sometimes referred to as d-c amplifiers. One of these uses a special tube known as the electrometer tube.[2] It is usually a one stage amplifier to be used with a galvanometer for measuring very small direct voltages.[3] In the other the d-c or low-frequency signal is modulated on a carrier of some higher frequency.[4] This carrier frequency is one that can be amplified by a conventional amplifier without the problems that arise when direct coupling is used. The modulated carrier can then be amplified and applied to a detector to give an output voltage which corresponds to the original signal input and is free of the carrier frequency. This type of circuit may be classed as a d-c amplifier but it is in no sense a direct-coupled amplifier.

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[1] Harold Goldberg, "A High Gain DC Amplifier for Bio-Electric Recording," Electrical Engineering, January, 1940, p. 60.

[2] W. B. Nottingham, "Measurement of Small DC Potentials and Currents in High Resistance Circuits by Using Vacuum Tubes," Journal of the Franklin Institute, March, 1930, p. 287.

[3] D. B. Penick, "Direct-Current Amplifier Circuits for Use With the Electrometer Tube," Review of Scientific Instruments, April, 1935, p. 115.

[4] L. J. Black and H. J. Scott, "A Direct-Current and Audio Frequency Amplifier," Proceedings of the I.R.E., June, 1940, p. 261.

IV. SCREEN-GRID COUPLING

The screen-grid coupled amplifier to be considered here is unique in that the signal is applied to the screen-grid of the vacuum tube rather than to the first, or control grid, as is usually the case. This use of the screen-grid greatly simplifies the problem of interstage coupling. The performance of the screen-grid coupled amplifier is similar to that of the more common d-c amplifiers. Simplicity is the principal advantage of the method.

CHAPTER II

REVIEW OF DIRECT-COUPLED AMPLIFIERS

The two principal problems arising when direct coupling is used are drift and interstage coupling. This chapter contains a brief review of the literature on compensated and balanced circuits for the reduction of drift and on the more common interstage coupling networks.

I. DRIFT

The principal problem introduced by direct coupling is that of drift, or a slow change in the output of an amplifier with no signal voltage applied at the input. In the resistance-capacitance-coupled or transformer-coupled amplifiers these small random variations of voltage may appear at the plate of each tube. However, since they are essentially d-c voltages, they will not appear at the grid of the following tube. Thus, while drift does occur in each individual tube, the effect is not amplified by following stages and does not appear in the output. In a direct-coupled amplifier these random variations of the plate voltage of one tube are amplified by succeeding stages. Thus a small drift in the first stage of an amplifier is likely to cause an objectionable change in the output.

BRIDGE BALANCE FOR SUPPLY VOLTAGE VARIATION

There are three primary causes of drift.[5] The first of these is variation of the plate-supply voltage. This drift may be remedied by complete regulation of the supply voltage. Another method of minimizing such drift uses the circuit shown in Fig. 1. Then, if the plate current is proportional to the supply voltage (i.e., R_b, the static plate resistance, is constant), the voltage between the plate of the tube and a point on the power supply bleeder is zero, independent of the supply voltage. This is a simple bridge circuit in which the equation for balance is

$$\frac{R_1}{R_2} = \frac{R}{R_b} . \qquad (1)$$

This balance holds true and the output is independent of the supply voltage over the range of voltages for which R_b is constant. The output voltage e_o is zero when no signal is applied. If the supply voltage E_b changes slightly, both I_p and I_1 change proportionally and e_o remains zero. The application of a signal voltage e_s to the grid causes I_p to change, thus changing the voltage developed across the load resistance R. There is no change in I_1. The output voltage is then equal to the voltage

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[5]Maurice Artzt, "Survey of D-C Amplifiers," Electronics, August, 1945, p. 112.

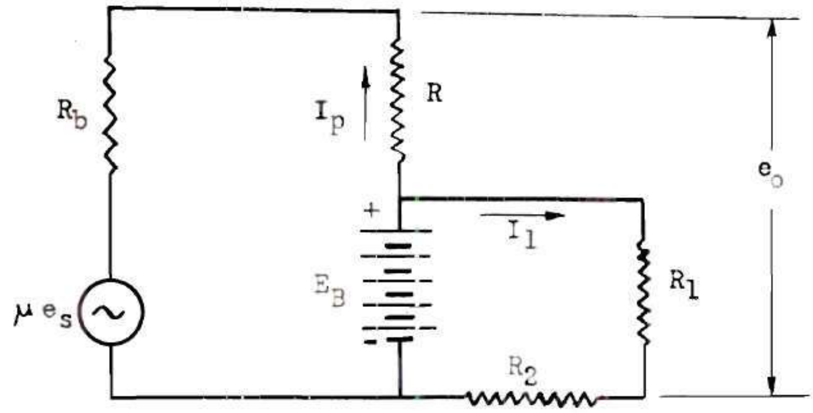

Fig. 1. Bridge Balance Circuit

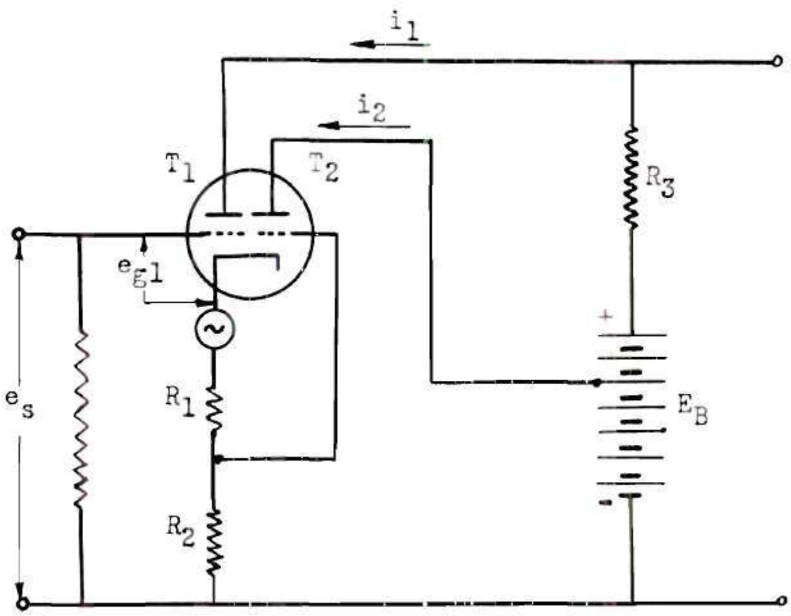

Fig. 2. Cathode Compensated Stage

developed across R due to the signal component of plate current i_p and is given by

$$e_o = \mu e_s \frac{R}{R + r_p} , \qquad (2)$$

where μ is the voltage amplification factor of the tube and r_p is the dynamic plate resistance. In the ideal case where perfect balance is to be achieved, R_b and r_p must be equal.

In the above discussion the internal impedance of the power supply has been considered to be zero. This, however, is not a necessary condition for balance. If the power supply impedance is considered, the bleeder current I_1 is not independent of the signal voltage and the output voltage of the power supply changes with load. However, as long as the bridge is balanced this variation of power supply voltage does not affect the output.

CATHODE COMPENSATION

The other two causes of drift are related and result from changes in cathode temperature. One of these causes is the change of contact potentials with cathode temperature and the other is the change of plate impedance with cathode temperature. A change of the contact potentials in the grid-cathode circuit changes the effective grid bias even though the externally applied bias remains unchanged. A change of the plate impedance results in a change in plate current with constant bias. Then, for the purpose of analysis, the effects of cathode temperature changes can be treated as due to changes in grid bias. This phenomenon is referred to as cathode drift.

The circuit of Fig. 2 can be made to compensate for this kind of drift.[6] The effective grid bias caused by cathode temperature effects is represented by the voltage v. One triode is used as the amplifier and the other for compensation only. It is desired to make i_1 independent of v so that the output voltage developed across R_3 will not be affected by small fluctuations in heater current.

The equations for the grid-cathode voltages of the two tubes of Fig. 2 are

$$e_{g1} = e_s + v - (i_1+i_2)\ (R_1+R_2) \quad , \tag{3}$$

$$e_{g2} = v - (i_1+i_2)R_1 \quad . \tag{4}$$

The incremental plate currents are given by[7]

$$i_1 = g_{m1}\ e_{g1} \quad , \tag{5}$$

$$i_2 = g_{m2}\ e_{g2} \quad . \tag{6}$$

Substituting equations (3) and (4) into these expressions for the currents we have

$$i_1 = g_{m1}\left[e_s + v - (i_1 + i_2)(R_1 + R_2)\right] \quad , \tag{7}$$

$$i_2 = g_{m2}\left[v - (i_1 + i_2)R_1\right] \quad . \tag{8}$$

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[6] Stewart E. Miller, "Sensitive D-C Amplifier With A-C Operation," Electronics, November, 1941, p. 27.

[7] The g_m used in equations (5) and (6) is the operating g_m of the tube measured along the dynamic load line.

These equations may be solved for the values of R_1 and R_2 necessary to make i_1 independent of v. This will be done by solving the equations for i_1 and then determining the element values necessary to make v drop out of the equation for i_1.

Solving equation (8) for i_2 we have

$$i_2 = \frac{g_{m2}(v - i_1R_1)}{1 + g_{m2}R_1} \quad . \tag{9}$$

Using this expression, the sum $i_1 + i_2$ is

$$i_1 + i_2 = \frac{i_1 + g_{m2}v}{1 + g_{m2}R_1} \quad . \tag{10}$$

Substituting this into equation (7) gives an expression for i_1.

$$i_1 = g_{m1} \; e_s + v - \frac{i_1 + g_{m2}v}{1 + g_{m2}R_1}(R_1 + R_2) \tag{11}$$

This may be rewritten

$$\frac{i_1}{g_{m1}} = e_s + v - \frac{i_1(R_1 + R_2)}{1 + g_{m2}R_1} - \frac{g_{m2}v}{1 + g_{m2}R_1}(R_1 + R_2) \; . \tag{12}$$

From equation (12) it is evident that i_1 will be independent of v if

$$\frac{g_{m2}(R_1 + R_2)}{1 + g_{m2}R_1} = 1. \tag{13}$$

This may be reduced to

$$g_{m2}R_1 + g_{m2}R_2 = 1 + g_{m2}R_1 \; , \tag{14}$$

or

$$g_{m2}R_2 = 1 \quad . \tag{15}$$

Equation (15) gives the condition necessary for compensation for cathode drift. A voltage v at the cathode causes a change of current in T_2 equal to $g_{m2}v$. This current flowing through R_2 causes a voltage drop equal to $g_{m2}R_2v$. This voltage drop is in phase opposition to v. If equation (15) is satisfied then $g_{m2}R_2v$ is equal to v. The voltage at the grid of T_1 is then the original drift voltage v in series with the compensating voltage -v. Since this sum is zero, the effective grid bias on T_1 is unchanged and its output is independent of small cathode drift voltages.

This compensation is seen to be independent of R_1. This resistance can therefore be adjusted for biasing as desired without upsetting the compensation for cathode drift. R_1 does not introduce any negative feedback since any change in the voltage across R_1 causes a compensating voltage to appear across R_2.

From the preceeding equations an expression may be written for the output voltage developed across R_3 and the gain of the stage shown to be[8]

$$\text{gain} = \frac{\mu R_3}{r_{p1} + R_3 + (\mu + 1)R_2} \quad . \tag{16}$$

Another method for reducing cathode drift is to provide a constant-current source for supplying heater current. This alone is not very satisfactory since it is impractical if not impossible to maintain heater current constant to the degree which would be required in a high-gain amplifier.

[8]See Appendix I.

II. INTERSTAGE COUPLING

Another problem encountered in the design of d-c amplifiers is that of interstage coupling. The output of one stage is taken from the plate of the tube and is therefore at some positive potential relative to its cathode. This voltage must be connected through a d-c path to the grid of the following tube. This grid is normally at some small negative potential relative to its cathode. It is therefore necessary either to reduce this positive potential before it is applied to the grid of the following tube or to raise the cathode potential of the second tube.

COUPLING THROUGH A BIASING BATTERY

The positive voltage may be reduced as much as is necessary by inserting a biasing battery in the grid circuit of the second tube.[9] This circuit, shown in Fig. 3, permits all cathodes to be grounded. The bias voltage on the grid of the second tube is then

$$E_{g2} = E_B - I_{p1}R - E_C \; . \qquad (16)$$

Then the entire signal voltage appearing at the plate of the first tube is applied to the following grid. A separate battery is required for the voltage E_C for each stage. Attempts to use a common E_C battery for all stages would result in all tubes so connected being parallel.

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[9] H. J. Reich, Theory and Applications of Electron Tubes (McGraw-Hill Book Company, Inc., 1939), p. 159.

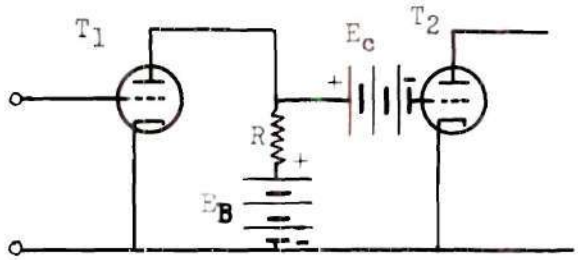

Fig. 3. Coupling Through a Biasing Battery

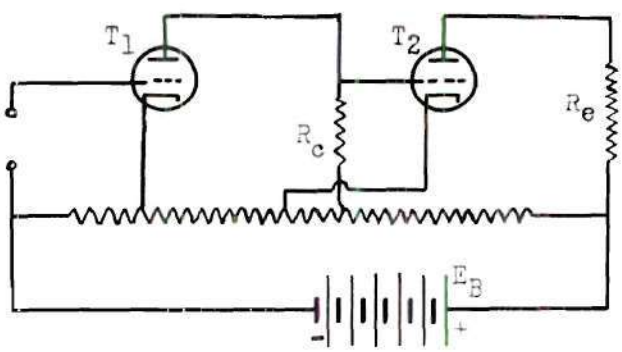

Fig. 4. Loftin-White Circuit

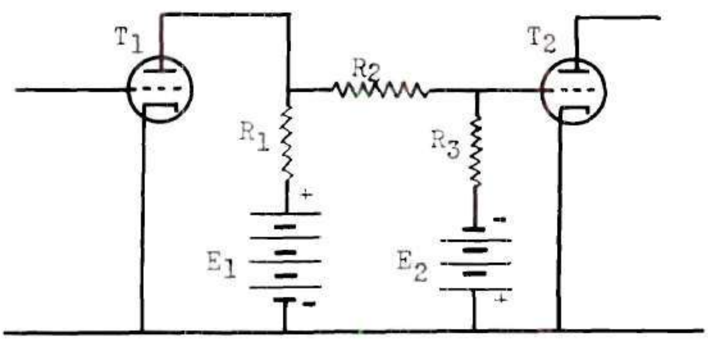

Fig. 5. Bridge Type Interstage Coupling

THE LOFTIN-WHITE CIRCUIT

A circuit in which the cathode potential of the second tube is raised[10] is shown in Fig. 4. Here successive stages are "stepped up" the power supply bleeder. The coupling resistor R_c is connected to the bleeder to give the desired plate voltage on the first tube. The cathode of the second tube is then connected to the point on the bleeder where the voltage is greater than the plate voltage of the first tube by an amount equal to the bias voltage desired on the grid of the second tube. The number of stages that can be cascaded in this manner is limited by the voltage of the power supply. The bleeder resistance must be such that the bleeder current will be large compared to the plate currents which flow. If this is not true, undesired coupling between stages will result.

BRIDGE-TYPE INTERSTAGE COUPLING

The bridge-type interstage coupling[11] of Fig. 5 permits all cathodes to be at the same potential but has the disadvantage of reducing the stage gain by the voltage division of R_2 and R_3. After the values of E_1 and E_2 have been fixed, R_1, R_2, and R_3 can be proportioned to give the desired values of plate voltage on the first tube, grid bias on the second tube, and load resistance in the plate circuit of the first tube. For example, using 6J7 tubes with 100 volts plate voltage, 100 volts screen voltage, -3 volts grid bias, and the suppressor connected to the cathode, the plate

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[10]Edward H. Loftin and S. Young White, "Cascaded Direct-Coupled Tube Systems Operated from Alternating Current," <u>Proceedings of the I.R.E.</u>, April, 1930, p. 669.

[11]W. M. Brubaker, "D-C Amplifier," <u>Bulletin of the American Physical Society</u>, December 7, 1939 (Program for the University of California meeting).

current will be 2 ma. If E_1 is 250 volts and E_2 is 150 volts, there must be voltage drops of 150 volts across R_1 and 250 volts across R_2 and R_3 in series. Since the grid bias on the second tube is to be -3 volts, there must be drops of 103 volts across R_2 and 147 volts across R_3. It is desirable to make the effective load resistance high in order to secure maximum gain. The value of R_1 is limited by the requirement that a current of approximately 2 ma produce an IR drop of 150 volts. If R_3 is made 1 megohm, R_2 must be 0.7 megohm. The current flowing through R_2 and R_3 will then be 147 μa. The total current flowing through R_1 is then approximately 2.15 ma. R_1 must then be 69,800 ohms. The effective load resistance in the plate circuit of the first tube is 68,200 ohms, and the amplification of the tube is 74.5. However, due to the voltage division across R_2 and R_3 this is reduced so that the gain of the entire stage is only 43.8. Only 58.8 per cent of the signal voltage at the plate of the first tube appears at the grid of the second tube. If the voltage E_2 can be increased, then R_3 can be increased and this loss of gain reduced. This circuit has the disadvantage of requiring very well regulated power supplies.

GAS-TUBE COUPLING

A similar circuit[12] which avoids the loss of gain due to the drop in R_2 is shown in Fig. 6. Here a gaseous diode is substituted for R_2. Since the voltage across such a tube is constant for a wide range of current, all of the signal voltage at the output of one stage appears at the grid of the following stage. The primary disadvantage of using gas-tube coupling is that gas tubes are inherently noisy.

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[12] Stewart E. Miller, op. cit.

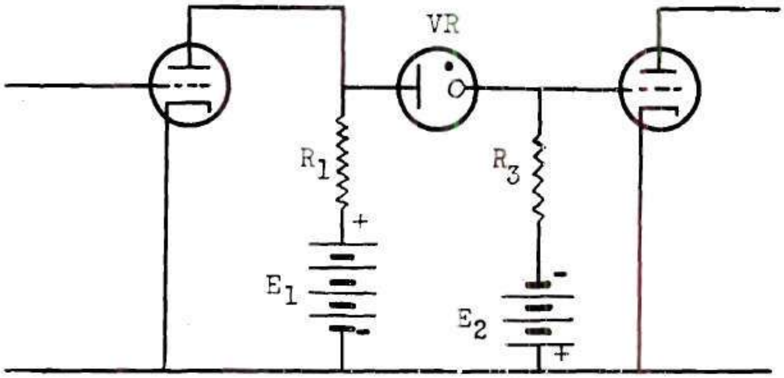

Fig. 6. Gas-Tube Coupling

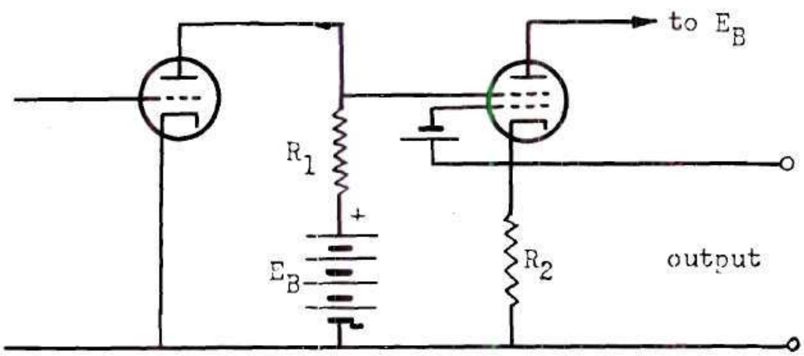

Fig. 7. Cathode Follower Coupling

CATHODE-FOLLOWER COUPLING

The circuit of Fig. 7 uses a screen-grid coupled cathode follower as an interstage coupling device.[13] The plate of the amplifier tube is directly connected to the screen grid of the following tube. The load resistance of this second tube is in the cathode circuit so that the output voltage developed across this resistance has one side at ground potential. It will then require only a small biasing voltage to provide a negative bias on the grid of the following amplifier tube. Since the entire output voltage is applied to the grid as a negative feedback voltage, the gain of the cathode follower stage is always less than unity and is commonly of the order of 0.7 to 0.8. This is true for both screen-grid coupled and control-grid coupled cathode follower circuits.

III. BALANCED D-C AMPLIFIER CIRCUITS

BASIC BRIDGE CIRCUIT

A basic type of balanced d-c amplifier[14] is shown in Fig. 8. This bridge circuit uses two similar tubes in opposite arms so that, when there is no signal, the bridge is balanced. If the two tubes are identical, the output voltage e_o is zero when e_s is zero for all values of supply voltage. Also, any change in the characteristics of one tube, if matched by an equal change in the other tube, does not unbalance the bridge

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[13]Y. P. Yu, "Cathode Follower Coupling in D-C Amplifiers," Electronics, August, 1946, p. 99.

[14]Maurice Artzt, op. cit.

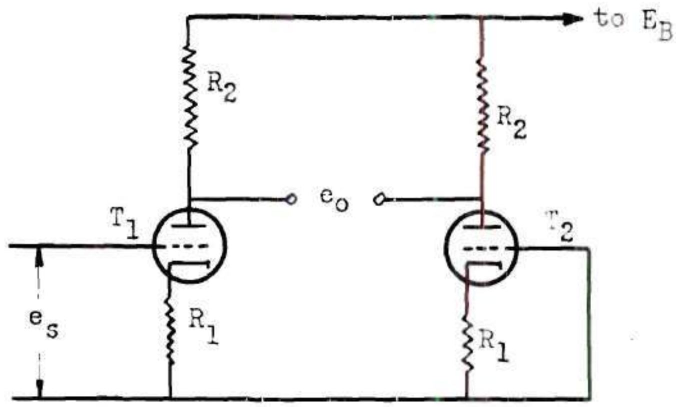

Fig. 8. Bridge-Balanced Amplifier

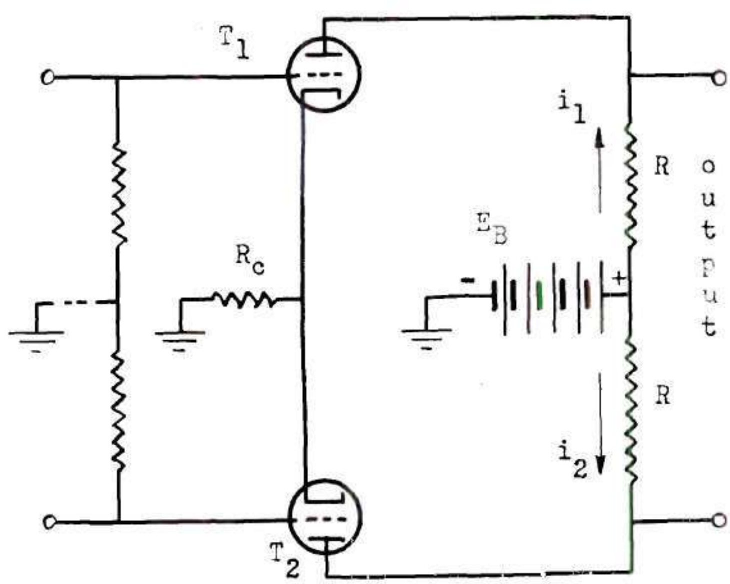

Fig. 9. Push-Pull Amplifier

and e_o remains zero. The application of a signal voltage e_s to the grid of only one tube unbalances the bridge and a voltage corresponding to the signal voltage appears at the output terminals.

Thus, if the two tubes remain identical, there is a perfect balance for variations in supply voltage and changes in tube characteristics. However, due to the normal differences in tubes, a perfect balance is rarely achieved. The circuit is, however, quite useful to minimize drift. Since the two tubes are normally mounted on the same chassis and draw their heater currents from the same source, any condition which tends to change the characteristics of one tube tends to have a similar effect on the other tube. The two changes then almost cancel each other and the bridge remains near balance.

There are several similar bridge-balanced d-c amplifier circuits to be found in the literature.[15] Such circuits prove effective in minimizing drift, especially if matched tubes can be used. A disadvantage of this type of circuit is that it is difficult to cascade several stages to increase the overall amplification.

PUSH-PULL CIRCUIT

The conventional push-pull amplifier circuit of Fig. 9 is basically the same as the bridge-balanced circuit of Fig. 8. As shown here there are only two differences. In the push-pull circuit the biasing resistor R_c is common to both tubes and does not cause the signal amplification to be reduced by negative feedback. Also, in the push-pull circuit, the signal is applied in opposite phase to both grids so that both tubes act as amplifiers.

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[15] Maurice Artzt, op. cit.

Characteristics of the Push-Pull Circuit

If the two tubes are operated on a linear part of their characteristic, a signal voltage at the input terminals produces equal but opposite changes in the plate currents. The sum of these currents, flowing through R_c and through the power supply, remains constant. Then the bias voltage developed across R_c remains constant and the signal amplification is not reduced by negative feedback. Since the current delivered by the power supply remains constant, undesired coupling between stages due to the impedance of a common power supply is also eliminated.

Phase Inversion

The push-pull amplifier gives a nearly balanced output even when the signal is applied to only one tube. In order for the output to be balanced it is necessary that $i_2 = -i_1$. If the signal is applied to the grid of tube 1 only it causes the plate current of tube 1 to change. This change in i_1 changes the voltage across R_c which is common to both tubes. Then the change in i_2 due to this change in bias voltage is given by

$$i_2 = -(i_1 + i_2)\, g_m R_c \quad . \qquad (17)$$

Solving this for i_2 we have

$$i_2 = -i_1 \frac{g_m R_c}{1 + g_m R_c} \quad . \qquad (18)$$

Then, if $g_m R_c$ is large compared to unity, i_2 is very nearly equal to the negative of i_1, and the output voltage is very nearly balanced.

The Cathode Resistor

In the push-pull amplifier of Fig. 9 it is possible to connect the plates of one stage directly to the grids of the following stage. The voltage of the cathode of the second tube can be raised by the use of a common cathode resistor to give the proper bias. Its resistance must be such that the sum of the two plate currents produces an IR drop equal to the desired bias voltage. As has been shown, this resistor will not introduce negative feedback if the circuit is balanced.

Illustrative Circuit

In illustration of this point the following data are given for a direct-coupled push-pull amplifier using 7F7 tubes and a 500 volt power supply. Suitable operation occurs with 99 volts plate voltage and -1 volt grid bias. The zero signal plate current per tube is 0.8 ma. Since the plate currents of both tubes flow through the biasing resistor, R_c for the first stage must be such that a current of 1.6 ma will develop a voltage of 1 volt. R_{c1} is then equal to 625 ohms. Since the cathode is now 1 volt above ground potential, a voltage of 99 volts across the tube puts the plate at 100 volts above ground. This requires that the plate current of 0.8 ma produce a voltage drop of 400 volts across the plate resistor R_1. R_1 must then be 500,000 ohms. This stage will have a voltage amplification of 70, or a gain of 36.9 db.

If the plates of the first stage are connected directly to the grids of the second stage, these grids will be at 100 volts above ground potential. The cathodes of the second stage must then be made 101 volts and the plates 200 volts above ground. This requires a cathode biasing resistor of 63,125 ohms and plate resistors of 375,000 ohms. The gain of this stage, 36.7 db, is slightly less than that of the first stage because of the smaller load resistance in the plate circuits of the second stage.

The third and fourth stages have gains of 36.3 db and 35.3 db respectively, giving an overall gain for four stages of 145.2 db or a voltage amplification of 1.82×10^7. The output of the fourth stage is 400 volts above ground potential.

IV. NEGATIVE FEEDBACK

Negative feedback can be used to reduce drift in a d-c amplifier just as it is used to stabilize ordinary amplifiers. If A is the gain of the amplifier and β is the fraction of the output voltage which is returned to the input, any drift introduced in the output stage is reduced by the factor $1 - A\beta$[16]. Drift in the input circuit is indistinguishable from the signal voltage and is not reduced. Drift introduced at points intermediate between the input and output is reduced to some extent, the reduction being greater when the point at which the drift is introduced is nearer the output.

Since most of the drift in the output of a d-c amplifier is due to drift of the initial stage, negative feedback would be of little value in minimizing this drift. However, if by some means the drift of the first stage can be minimized, negative feedback from the output of the second stage to the grid of the first stage would reduce the drift introduced in the second stage. Negative feedback is of no value for reducing drift when used alone. When used in conjunction with a suitable

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[16] F. E. Terman, "Feedback Amplifier Design," Electronics, January, 1937, p. 12.

drift reducing circuit such as the cathode compensation circuit of Fig. 2 it does prove to be advantageous. In any case, feedback will stabilize the gain of the amplifier. With some arrangement such as that just described it can be utilized to aid in the reduction of drift.

CHAPTER III

THEORY OF SCREEN-GRID COUPLING

I. THE SCREEN-GRID TUBE

A screen-grid tube has in addition to the plate, cathode, and control grid of the triode, a second grid called the screen grid. The screen grid is located between the control grid and the plate and is normally operated at a fixed positive potential. It constitutes a nearly perfect electrostatic shield between the control grid and the plate. It therefore assumes control of the plate current. Since this grid is at a positive potential and is located between the cathode and the plate, some of the electrons emitted from the cathode will go to the screen grid. However, most of the electrons drawn from the cathode will pass through the screen-grid mesh and go on to the plate. The screen grid then has a considerable effect on the plate current but intercepts only a relatively small proportion of the total electron current.

II. THE PROBLEM OF INTERSTAGE COUPLING

In the conventional amplifier the interstage-coupling network contains a blocking condenser which serves to isolate the positive plate voltage of one stage from the negative grid bias of the following stage. In a direct-coupled amplifier blocking condensers cannot be used. The problem therefore arises of coupling the plate of one stage to the grid of the following stage through a d-c path. Several circuits for satisfactory interstage coupling have been described in Chapter II.

The problem of interstage coupling in d-c amplifiers would be greatly simplified if the output of one stage could be applied to a grid of the following stage which is normally at a positibe potential comparable to that of the plate of the first stage. The screen grid of a tetrode or pentode is such a grid. It is usually operated at a positive potential comparable and commonly equal to the plate voltage.

Coupling directly from the plate of one stage to the screen grid of the following stage presents no serious problem. Such an arrangement can be made to give reasonable voltage amplification and a very simple amplifier results. It is related to the screen-grid coupled cathode follower stage shown in Fig. 7. There, however, no attempt was made to get any voltage amplification in the screen-grid coupled stage.

III. TUBE FACTORS FOR THE SCREEN GRID

In order to study the performance of the screen-grid coupled amplifier, the usual tube factors (μ, g_m, and r_p) must be supplemented by similar factors applicable to the screen grid. Since the screen-grid circuit carries a current which is a function of the screen voltage, the dynamic screen-grid resistance may be defined as

$$r_{sg} = \frac{1}{\dfrac{\partial i_{sg}}{\partial e_{sg}}} \Bigg|_{e_{g1}, e_p \text{ constant.}} \qquad (1)$$

The reciprocal form is used in order to comply with the convention of putting the independent variable in the denominator of the derivative.

It has been found convenient to use the ratio of incremental plate current to incremental screen current in some of the calculations. The symbol μ_i is considered appropriate for this ratio because μ is commonly used to represent a dimensionless amplification ratio. The subscript i is used to distinquish it from any voltage amplification ratio. The dynamic screen-plate current amplification ratio μ_i may then de defined as

$$\mu_i = \left.\frac{\partial i_p}{\partial i_{sg}}\right|_{e_{g1},\ e_p \text{ constant}} . \qquad (2)$$

The screen-plate transconductance may be defined as

$$g_{m2} = \left.\frac{\partial i_p}{\partial e_{sg}}\right|_{e_{g1},\ e_p \text{ constant}} . \qquad (3)$$

It is evident from equations (1), (2), and (3) that

$$g_{m2}\, r_{sg} = \mu_i \quad . \qquad (4)$$

It is sometimes inconvenient to calculate one of these factors directly from the data available. In such a case it can usually be evaluated by using equation (4).

The dynamic plate resistance may be defined as

$$r_p = \left.\frac{1}{\frac{\partial i_p}{\partial e_p}}\right|_{e_{g1},\ e_{sg} \text{ constant}} . \qquad (5)$$

With the four factors defined in equations (1), (2), (3), and (5) the performance of the screen-grid coupled amplifier stage may be analyzed.

IV. GAIN OF A SINGLE STAGE

Using the above factors an expression for the gain of a screen-grid coupled stage may now be derived. Assuming the control grid voltage constant, the incremental plate current is

$$i_p = \frac{\partial i_p}{\partial e_{sg}} e_{sg} + \frac{\partial i_p}{\partial e_p} e_p \quad . \tag{6}$$

In this equation and in those which follow, lower-case letters represent incremental values. Substituting the tube factors defined in equations (3) and (5), equation (6) becomes

$$i_p = g_{m2}\, e_{sg} + \frac{e_p}{r_p} \quad . \tag{7}$$

The actual instantaneous plate voltage is the constant plate-supply voltage less the voltage drop caused by the plate current flowing through the load resistor. Thus the incremental plate voltage is the negative of this i_pR[17] drop, or

$$e_p = -i_pR \quad . \tag{8}$$

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[17]The load resistance R used in these equations is the effective load resistance in the plate circuit of the tube. In a circuit where the plate of the tube under consideration is connected directly to the screen of another tube, this effective load resistance will include the dynamic screen resistance of the following tube.

Making this substitution in equation (7) we have

$$i_p = g_{m2}e_{sg} - i_p \frac{R}{r_p} \quad . \tag{9}$$

Equation (9) can be solved for i_p to give

$$i_p = g_{m2}\, e_{sg} \frac{r_p}{R + r_p} \quad . \tag{10}$$

The output voltage is developed by this plate current flowing through the load resistance R and is given by

$$e_o = i_p R = g_{m2}\, e_{sg} \frac{Rr_p}{R + r_p} \quad . \tag{11}$$

The stage gain is then

$$\frac{e_o}{e_{sg}} = g_{m2} \frac{Rr_p}{R + r_p} \quad . \tag{12}$$

This may be recognized as the same expression for gain as might be derived from an equivalent circuit of the constant-current generator form.

V. CONDITIONS FOR MAXIMUM GAIN

In order to determine the tube characteristics desired for maximum stage gain consider a simple stage such as is shown in Fig. 10. The effective load resistance in the plate circuit of T_1 is the coupling resistance R_c in parallel with the dynamic screen resistance of T_2.

$$R = \frac{R_c r_{sg}}{R_c + r_{sg}} \tag{13}$$

Making this substitution, equation (12) for the gain of one stage becomes

$$\text{gain} = g_{m2} \frac{1}{\frac{1}{r_{sg}} + \frac{1}{r_p} + \frac{1}{R_c}} \quad . \tag{14}$$

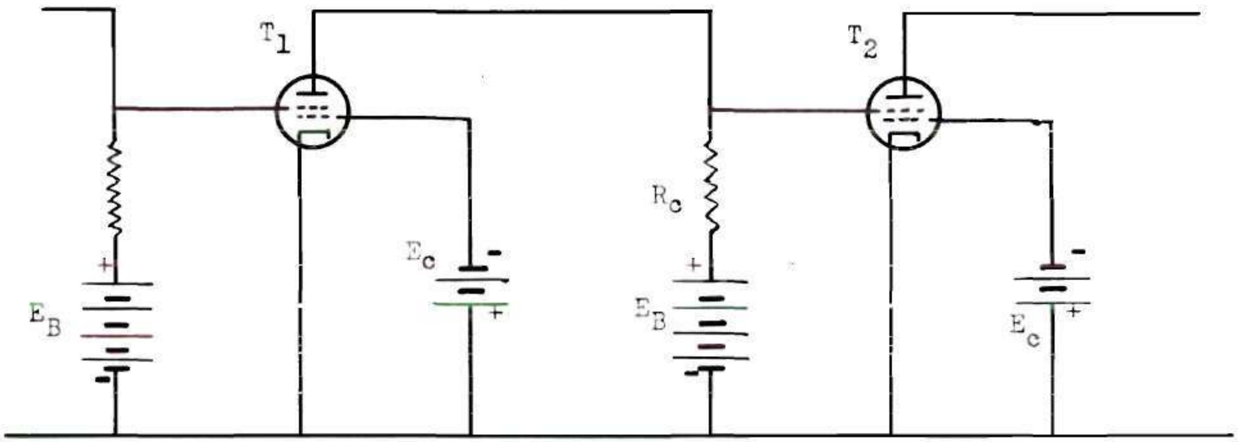

Fig. 10. A Typical Screen-Grid Coupled Stage

Substituting for g_{m2} from equation (4) this becomes

$$\text{gain} = \mu_i \frac{1}{1 + \frac{r_{sg}}{r_p} + \frac{r_{sg}}{R_c}} \quad . \tag{15}$$

From equation (15) it can be seen that the gain is proportional to μ_i and cannot be larger than μ_i. It would then be desirable to use a tube with a large ratio of plate to screen current at the operating point. Several of the more common beam pentodes have current ratios in excess of 20. It is also desirable that r_{sg} be small and that r_p and R_c be as large as possible. Since R_c is an actual physical resistor, its value can be adjusted as necessary. The various factors considered in choosing R_c will be discussed in connection with the design of the amplifier in a later section. The other three parameters, μ_i, r_p, and r_{sg}, are tube characteristics. It is therefore not possible to adjust one of them individually without influencing the others.

It should be noted that r_p is the plate resistance of T_1 and that r_{sg} is the screen resistance of the following tube T_2. However, the two tubes would normally have comparable operating points so that the screen resistance of the two tubes will be approximately equal.

CHAPTER IV

DESIGN OF THE AMPLIFIER

I. SELECTION OF TUBES

The first problem encountered in the design of the amplifier is the selection of tubes. The tube characteristics desired for maximum gain are a high ratio of plate current to screen current, a high dynamic plate resistance, and a low dynamic screen resistance. It would also be desirable to use a tube which requires low filament power and one which is common enough to be readily available. However, the primary considerations are those concerned with maximum gain.

TYPICAL PENTODE CHARACTERISTICS

Fig. 11 shows a typical pentode characteristic curve of plate current versus plate voltage with screen voltage and control-grid voltage held constant. In these coordinates a horizontal line represents an infinite resistance. The pentode has a high dynamic plate resistance in the normal operating range. In the screen-grid coupled amplifier, the tubes must be operated with plate and screen voltages equal, or approximately equal. In Fig. 11 the plate and screen voltages are equal at the point marked A. This point is beyond the knee of the curve or in the range of high dynamic plate resistance. However, since it is not far beyond the knee of the curve, the magnitude of signal voltages that can be passed without non-linear amplification of the negative peaks is limited.

The dynamic screen resistance of a pentode is also high due to the small values of screen current. A change in screen voltage may cause a large per cent change in current, but, since the resistance is determined by the actual magnitude of the change rather than the relative change, the resistance will be large.

The current amplification ratio varies from 2 to 25 for different types of commercial pentodes. There is also considerable variation of this ratio in different tubes of the same type. Tube manufacturers attempt to make the screen current low in order to reduce the power consumed by the screen grid, which normally has no active part in the operation

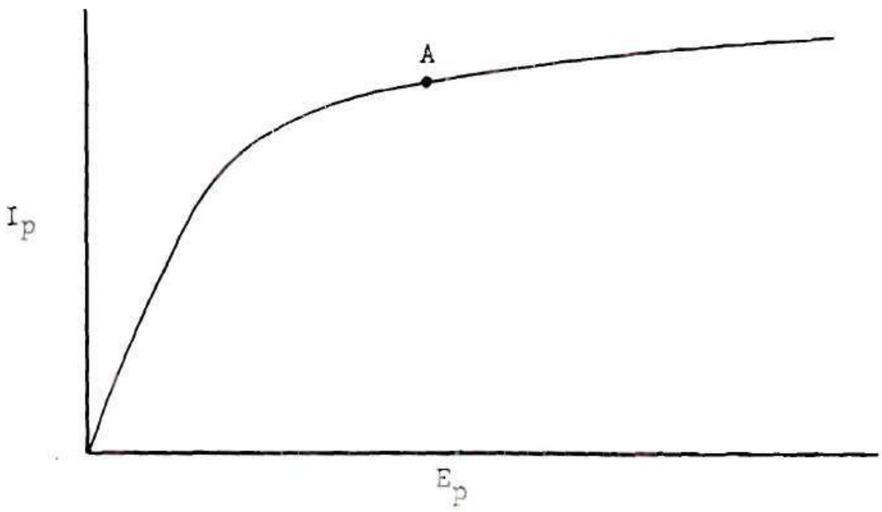

Fig. 11. Typical Pentode Characteristic

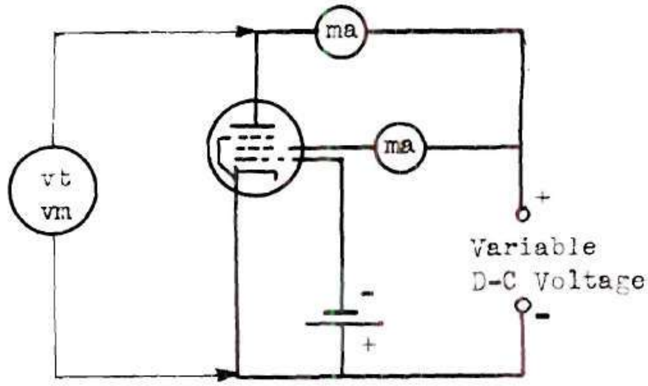

Fig. 12. Circuit For Testing Pentode Current Amplification

of the tube. However, this current ratio is normally a secondary characteristic, so not as much effort is expended in making it consistent as is devoted to other more common characteristics. The result is that, even in supposedly similar tubes, the current amplification μ_i is rather variable. There are, however, a few types which show a reasonable consistency of this characteristic. A satisfactory amplifier could be built with randomly selected tubes, but better performance is had with picked tubes.

BEAM POWER TUBES

Beam power tubes in general have higher values of μ_i than ordinary pentodes. A beam power tube is a tetrode or pentode in which use is made of directed electron beams to increase the power handling capacity of the tube. Such tubes contain a cathode, control grid, screen grid, plate, and beam forming electrodes. These beam-forming plates are at cathode potential and assist in producing the desired beam effects. Secondary emission current from the plate is suppressed by space-charge effects between the screen and plate. The screen and control grid are spiral wires so wound that each turn of the screen is shaded from the cathode by a grid turn. This alignment of the screen and control grids causes the electrons to flow in sheets between the turns of the screen so that few of them reach the screen. Thus, a feature of the beam power tube is low screen current or a high ratio of plate current to screen current.

Since the plate is more effectively shielded from the cathode than in the ordinary pentode, the plate resistance of the beam power tube is relatively high. The small values of screen current result in high screen resistances.

TESTS MADE FOR SELECTING TUBES

A number of tubes, both beam power tubes and ordinary pentodes, were tested in an effort to find tubes with a high current amplification. The only information of interest normally given in the manufacturers tube manuals is the zero-signal plate and screen currents for one or two specific operating points. There is sometimes a curve of screen current versus plate voltage on the plate characteristic curve sheet. From this kind of data some tubes were selected for tests. Most of the tubes tested were taken from stocks on hand in order to obtain data on a variety of tubes. When available, at least two tubes of each type were tested.

Preliminary tests were made using the circuit shown in Fig. 12. With plate and screen voltages equal, plate current and screen current were measured for several values of control-grid bias. With constant control-grid bias, the plate current was varied from nearly zero to values higher than normal. The tests were repeated for several values of control grid bias voltage.

RESULTS OF TESTS

Typical results of these tests are shown in Table I. The three tubes marked with an asterisk are beam power tubes. The others are ordinary pentodes. The values shown for μ_i are the highest values found to be consistent over an appreciable range of plate voltage variation and are an average of this value for all tubes of the same type tested.

The beam power tubes have a current amplification of more than twice that of any of the ordinary pentodes tested. The plate resistance of the beam power tubes is higher so as to further increase the gain. The screen resistance is also higher, but this disadvantage is more than offset by the much higher value of μ_i.

In the beam power tubes, the highest current amplification occurs with currents much smaller than the usual values. For example, the 6V6, frequently operated with plate currents of 30 ma and higher, shows the highest current amplification when the plate current is less than 1 ma.

TABLE I

Tubes Tested for High μ_i

Tube Type	Number Tested	Average μ_i
1S4	2	5
1S5	2	5
1T4	2	3
1T5*	3	20
3A4	2	8
3S4	1	5
6AG5	1	3
6AJ5	2	3
6AK5	3	3
6AK6	3	6
6BA6	1	3
6L6*	2	25
6V6*	2	25
9001	2	3
9003	2	3

*Beam power tubes

Further tests were made on some of the tubes, particularly the 6AK6, to determine the effect of operating the suppressor at various positive and negative biases. As the suppressor voltage is made negative, the plate current decreases and the screen current increases. When the suppressor is connected to the plate, the plate-to-screen current ratio increases slightly but is still much lower than the ratios found in beam power tubes.

Tests with reduced heater voltage showed only slight changes in current amplification for the same current magnitudes. The principal result of reducing the heater voltage is that higher plate and screen voltages are required to maintain the currents constant as the heater voltage is reduced. In the 6V6 no appreciable change in current amplification occured over the range 3.0 to 6.3 volts.

Two tubes, the 6V6 and the 1T5, were selected for more extensive tests. The 6L6 showed characteristics very similar to the 6V6. The 6V6 was favored because it requires only half as much filament power as does the 6L6. The test data indicates that a satisfactory amplifier could be built using either of these tree tubes. The 6V6 was chosen over the 1T5 because it showed slightly superior characteristics. The 6V6 is also a more common tube and is known to be very rugged.

II. THE 6V6 STAGE

NECESSARY CURVES AND DATA

Two sets of characteristic curves were drawn for the 6V6. The first of these is given in Fig. 20, Appendix II. These curves, taken with plate and screen voltages equal, show screen current as a function of this voltage. Five such curves are plotted, each for a different control-grid bias voltage. On the same sheet are plotted another family of curves showing the current amplification for the same control-grid bias voltages and over the same range of plate and screen voltages. From these curves the current amplification and approximate magnitudes of plate and screen currents can be determined for various combinations of plate voltage and bias voltage. Conditions for high current amplification can be very easily seen.

A second family of curves is given in Figs. 21, 22, and 23. Each curve sheet contains several curves taken with the same constant control-grid bias voltage. Each curve on the sheet is taken with screen voltage constant. The curves then show plate current versus plate voltage and screen current versus plate voltage for constant values of screen and control-grid voltage. It is from these curves that the tube factors defined in Chapter III may be calculated.

Further data on a typical 6V6 are given in Table IV. Here are tabulated various combinations of plate, screen, and control-grid voltages necessary to give plate currents of approximately 0.85 to 0.90 ma. Screen current for the various conditions is also tabulated. This data can be used to supplement that given in the curves of Appendix II.

CALCULATION OF TUBE FACTORS FROM CURVES

To illustrate the calculation of tube factors from the curves we may choose a typical operating point on one of the curve sheets of Appendix II. With control grid bias of -7.5 volts, screen voltage of 60 volts, and plate voltage of 90 volts, the plate current is 0.7 ma and the screen current is 22.5 μa. The current amplification is then given by[18]

$$\mu_i = \frac{700}{22.5} = 31 \quad .$$

The plate resistance is defined by the slope of the plate current curve at the operating point. The plate resistance r_p is then approximately 625,000 ohms.

The screen resistance may be calculated from the change in screen current corresponding to an appropriate change of screen voltage along

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[18]This value is only an approximate one as will be shown later in this section.

the vertical line representing constant plate voltage. In this case the appropriate screen voltage increment would be from 55 volts to 65 volts, or 5 volts above and below the operating point. With the plate voltage equal to 90 volts and the screen voltage 55 volts, the screen current is 15.5 μa. When the screen voltage is increased to 65 volts the screen current will increase to 31 μa. The screen resistance is then

$$r_{sg} = \frac{\Delta e_{sg}}{\Delta i_{sg}} = \frac{10}{15.5} \times 10^6 \text{ ohms} = 645{,}000 \text{ ohms} .$$

The transconductance may be calculated in a similar manner from the change in plate current brought about by this same screen-voltage increment. With the plate voltage constant at 90 volts, increasing the screen voltage from 55 to 65 volts causes the plate current to increase from 0.45 ma to 0.99 ma. The transconductance is therefore

$$g_{m2} = \frac{0.54 \text{ ma}}{10 \text{ v}} = 54 \times 10^{-6} \text{ mhos} .$$

From equation (4) in Chapter III it is seen that

$$\mu_i = g_{m2} r_{sg} .$$

Since g_{m2} and r_{sg} are now known, this relation may be used to determine μ_i. Making the substitution we have

$$\mu_i = 54 \times 10^{-6} \times 645{,}000 = 34.8$$

This is slightly larger than the value calculated from the ratio of plate current to screen current at the operating point. This discrepancy may be explained as follows. The quantity μ_i as defined and used in Chapter III is the dynamic current amplification, or the change in plate current corresponding to a small change in screen current. The

μ_i calculated from the curves to be 31 is the simple static ratio of the total plate current to the total screen current at the operating point. This static ratio is only an approximation of the actual current amplification. Comparison of the two values calculated here shows that the static current ratio does give a reasonably close approximation.

SELECTION OF AN OPERATING POINT

The two primary considerations in the selection of the operating point are gain and interstage coupling. The current amplification at the operating point should be relatively high. The screen voltage of the particular tube under consideration will be the same as the plate voltage of the preceding tube. Also, the plate voltage will be the same as the screen voltage of the following tube. It is therefore impractical to consider the operating point of one tube alone. A design of the complete circuit must be worked out so that all tubes will have reasonable gain and voltages suitable for coupling to the adjacent tubes.

Plate and Screen Voltages Equal

The simplest multistage amplifier is one in which all stages are identical - i.e., have the same voltages, the same load resistance, etc. Each stage therefore has its plate voltage equal to its screen voltage. The optimum current amplification for this condition is of the order of 20 to 25. As has been previously pointed out, the point at which the plate and screen voltages are equal is near the knee of the plate current vs. plate voltage curve (see Fig. 11) so that the magnitude of signal voltage which can be passed without distortion is uncomfortably small.

A further disadvantage of this design is that, due to the normal differences in tubes, it is virtually impossible to have all stages identical. However, the plate and screen voltages can be made very nearly the same by providing a control-grid bias adjustment on each stage.

If in any stage the plate voltage is reduced to a value only a few volts below the screen voltage, the plate current drops and the screen current rises sharply. Thus the current amplification is greatly reduced. Any drift of voltage in the initial stages is therefore likely to cause the gain of a later stage to be reduced.

Plate Voltage Higher than Screen Voltage

If the plate voltage of the tube is raised while its screen voltage remains constant, the plate current will increase and the screen current will decrease. Thus the current amplification increases. With the screen voltage equal to 60 volts and control grid bias of -7.5 volts, increasing the plate voltage from 60 to 90 volts increases the static current ratio from 22 to 31. Thus the gain of such a stage could be increased considerably. Increasing the plate voltage also increases the magnitude of signal that can be passed without distortion.

Cascading several stages is somewhat more difficult when the plate voltages are higher than their respective screen voltages than if all voltages are equal. This is especially true if the available power-supply voltage is limited, because each plate is necessarily at a higher voltage than the preceding plate. This difficulty will be discussed in more detail later.

THE LOAD RESISTOR

General Considerations

The load resistor must be of such a magnitude that the sum of the plate current of one tube and the screen current of the following tube flowing through this resistor will produce a voltage drop equal to the difference between the supply voltage and the plate voltage. It is desirable to make this resistance high in order to have a high gain. Use of a higher resistance decreases the bandwidth by reducing the high frequency cut-off. Moreover, the magnitude of the load resistor is limited by the voltage of the power supply. As the resistance is increased, the supply voltage must also be increased to maintain the plate voltage at the desired value.

As an example, let us calculate the value of the load resistor for a stage with the operating point defined for the calculation of tube coefficients and with a supply voltage of 300 volts. The plate voltage at this point is 90 volts. The voltage drop across the load resistor must therefore be 210 volts. This voltage drop is produced by the plate current, which is 0.7 ma, and the screen current of the following tube, which is not yet known.

It is therefore necessary to fix the operating point of the following stage before continuing. Its screen voltage is known to be 90 volts. Again, the plate voltage can be made 30 volts higher than the screen voltage, or 120 volts. Since the higher current amplification occurs with low magnitudes of current, the control-grid bias voltage will be made such that approximately the same currents flow in this stage as in the previous stage. This, of course, does not specifically define the operating point but it is sufficient for the present purpose. The screen current will be approximately 25 μa.

The current in the load resistor is then 0.725 ma. The resistance is given by

$$R = \frac{210}{0.725} \times 10^3 = 290{,}000 \text{ ohms} \quad .$$

If a suitable resistor of this value is not available, one of a slightly different value (300,000 ohms, for instance) may be substituted. In either case, after the resistor has been connected in the circuit, the control-grid bias voltage is adjusted until the proper voltage appears at the plate.

The plate voltages are much more critical in a direct-coupled amplifier than in other amplifiers. To illustrate this assume a uniform stage amplification of 10. If the plate voltage of one stage is changed by one volt (possibly by changing the tube) the screen voltage of the following tube is changed by the same amount. The plate voltage of this second tube is then changed 10 volts and that of the third tube 100 volts. The plate voltages must therefore be kept very close to the design values or the following tubes will be far off their proper operating points.

The curves of Fig. 13 show stage gain plotted against load resistance for three different conditions. These curves are based on calculations using data obtained from Appendix II. The plate current is constant at 0.8 ma for all of the curves. Each curve represents a different combination of constant plate and screen voltages. The power-supply voltage necessary for these same conditions is also plotted against load resistance on the same sheet.

These curves are not intended to give accurately the gain available from a 6V6 stage operating under the conditions shown. Due to the differences that normally exist between tubes (especially of a secondary characteristic such as is being exploited here) it would be most unusual

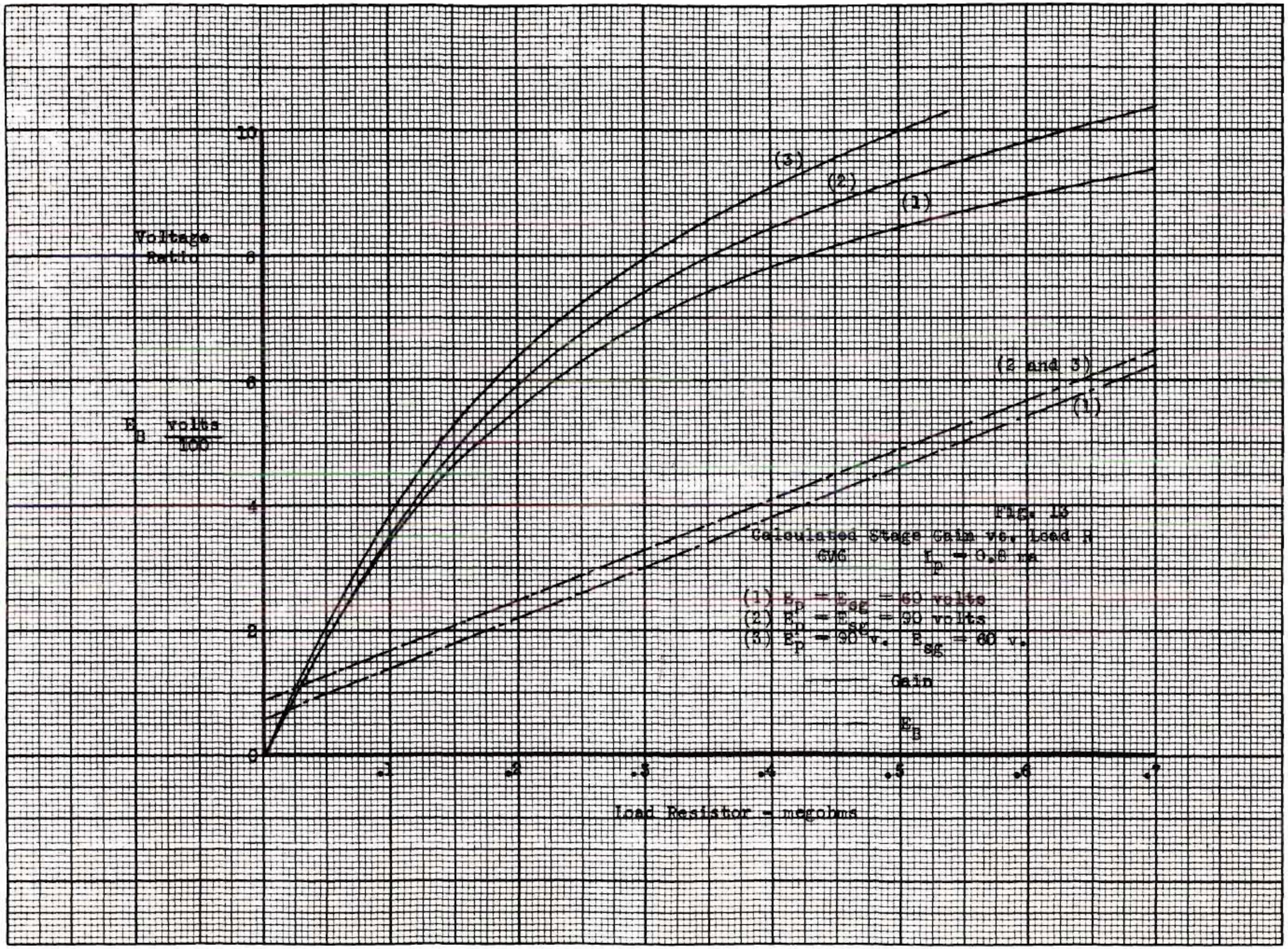


Fig. 13
Calculated Stage Gain vs. Load R
6V6 I_p = 0.8 ma

to find two tubes which show exactly the same gain characteristics. The curves do, however, show the manner in which gain varies with load resistance for a typical tube. The performance of most tubes would be comparable even though there may be appreciable differences in actual magnitudes.

It is evident from the curves that increasing the load resistance will increase the stage gain. However, since the plate voltage and plate current are constant, increasing the load resistance will increase the voltage required from the power supply. It is therefore necessary to compromise in order to secure a reasonably high gain with a reasonably low power supply voltage. For the conditions shown on Fig. 13, satisfactory results would probably be had with a load resistance of approximately 0.25 megohm and a 300 volt power supply.

The Dynamic Load Resistance

The dynamic load resistance of any stage consists of the load resistor in parallel with the screen resistance of the following tube. For the conditions already described, the screen resistance is approximately 650,000 ohms. If the load resistor is 300,000 ohms, the dynamic load resistance will then be

$$R = \frac{650{,}000 \times 300{,}000}{950{,}000} = 205{,}000 \text{ ohms} \quad .$$

Gain of One Stage

The gain of the complete stage shown in Fig. 14 may now be calculated. From equation (12), Chapter III, we have

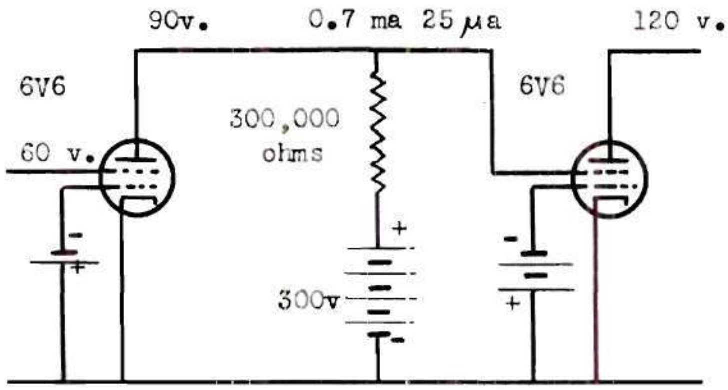


Fig. 14. A Typical 6V6 Stage

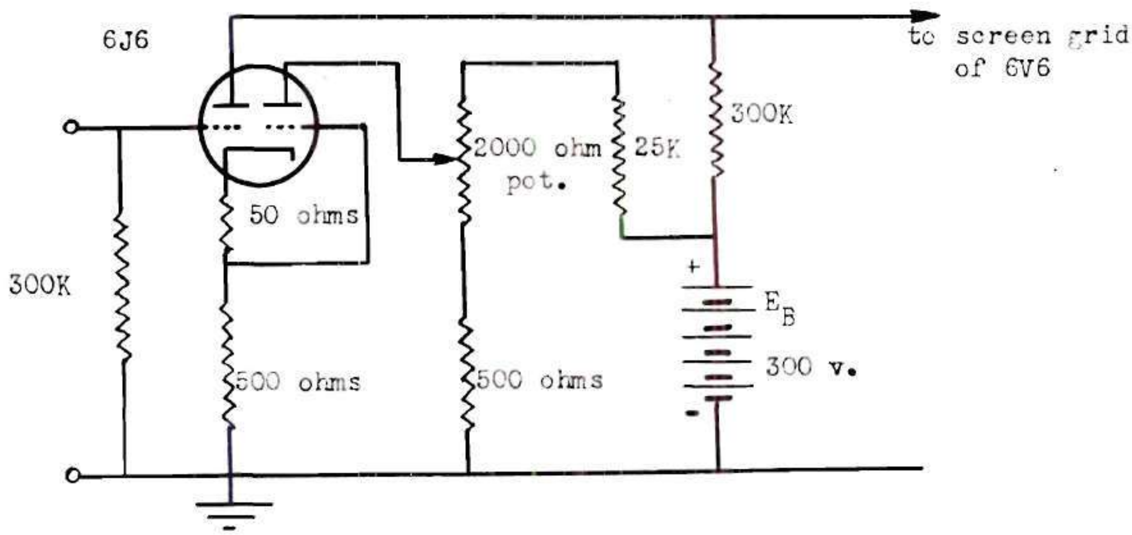


Fig. 15. 6J6 Input Stage

$$\text{gain} = g_{m2} \frac{R\, r_p}{R + r_p} \quad .$$

The values of g_{m2} and r_p have been determined and are 54×10^{-6} mhos and 625,000 ohms respectively. The stage gain is then

$$\text{gain} = 54 \times 10^{-6} \frac{205{,}000 \times 625{,}000}{205{,}000 + 625{,}000} = 8.4 \quad .$$

It was necessary to use wire-wound resistors in the first four stages. When carbon resistors were used in the initial stages the drift was much worse than when wire-wound resistors were used. Carbon resistors are inherently noisy. Their use in a d-c amplifier results in a large and somewhat erratic drift that is intolerable. Carbon resistors could be safely used in the control-grid filter circuit since that resistor carried no current. Wire-wound resistors were used in all of the bias-supply bleeders and in the plate circuits of the first four stages.

III. INPUT CIRCUIT

The input to the first stage of the amplifier need not be applied to the screen grid. The purpose of using the screen grid in later stages is to avoid the necessity of reducing the positive voltage taken from the plate of one stage to a small negative voltage suitable for application to the control grid of the following tube. Since the input is not at such a positive voltage it may be applied directly to the control grid. In fact this will be considerably simpler than superimposing the signal on a positive voltage suitable for application to the screen grid.

CATHODE COMPENSATION

The first stage should employ some arrangement for minimizing drift. Since a triode input may be used, the cathode compensation circuit of Fig. 2 appears attractive. This circuit makes use of the effects of the cathode drift voltage of one triode unit to compensate for the effects of this same voltage on the other triode. It is therefore highly desirable to use a dual triode tube, preferably one in which the same physical cathode structure serves both triode units. If there is actually a common cathode rather than two similar cathodes connected together inside the glass envelope, then the cathode drift voltage of the two units will be more nearly identical and the compensation will be more effective.

The 6J6, a miniature dual triode, has a common cathode for both triode units. The electrical characteristics of each triode are similar to those of the 6J5.

DESIGN AND ADJUSTMENT

The condition for compensation is that the cathode resistor (R_2) be equal to the reciprocal of the operating g_m of the compensating triode. Further considerations in designing the stage are that the plate voltage of the amplifying triode should be kept low so that it will be suitable for coupling to the screen of the following tube, and the plate currents should be made as low as practical. Low plate currents are desirable because they permit the use of a higher load resistor and thus increase the gain and because high currents would produce an excessive bias voltage across the compensating resistor R_2.

The final design of the input stage is shown in Fig. 15. Approximate values for the various elements were arrived at after a study of the

6J5 characteristics. An experimental circuit was constructed and the circuit elements adjusted to give optimum performance. Adjustments were made with a small 1000-cycle signal applied across the 50 ohm cathode resistor. The 2000 ohm potentiometer in the plate circuit of the compensating triode was then adjusted until the 1000-cycle output voltage across the 300,000 ohm load resistor was reduced to zero. This adjustment must be made each time the tube is replaced. No readjustment is needed as long as the same tube remains in the circuit.

The compensation is most satisfactory when the plate voltage of the compensating triode is approximately 23 volts and that of the amplifying triode is 50 volts. In this condition the amplification factor of the amplifying triode is approximately 17 and the plate resistance 15,000 ohms. The gain may then be calculated to be

$$\text{gain} = 17\,\frac{300{,}000}{315{,}000} = 16 \quad .$$

The measured gain of the experimental circuit was 16 at 1000 cycles.

IV. THE COMPLETE AMPLIFIER

THE OPERATING POINTS

The use of plate voltages higher than the corresponding screen voltages is considered advantageous because it makes possible higher stage gains and larger undistorted signals. The principal disadvantage of this arrangement is that it may require high supply voltages on the latter stages.

In order to have a uniform stage gain the plate voltages of the various stages should exceed their respective screen voltages by approximately the same ratio rather than by the same number of volts. This can be seen by referring to the curves of Figs. 21, 22, and 23 in Appendix II. Since all three curve sheets have the same scales it is permissible to use linear dimensions as a basis for comparison. With comparable currents, the vertical distance between corresponding plate and screen current curves is less for the higher values of screen voltage. Then with the greater screen voltage a proportionally greater plate voltage is required to give the same current ratio.

For example, compare the 65 volt curves of Fig. 21, the 75 volt curves of Fig. 22, and the 85 volt curves of Fig. 23. The scales are such that the ratio of currents is twenty at the point of intersection. On Fig. 21 the plate voltage must be 85, or 20 volts higher than the screen voltage to give a current ratio of thirty. The increase of plate voltage required for this same increase of current ratio is 25 volts for Fig. 22 and 30 volts for Fig. 23. The ratio of plate voltage to screen voltage for these three conditions is 1.31, 1.33, and 1.35 respectively.

The Second Stage

The output of the first stage is taken from the plate of the triode which is 50 volts above ground potential. The screen voltage of the 6V6 in the second stage is then 50 volts. Let the plate voltage of the second stage be 60 volts and the grid bias such that the plate current will be 0.85 ma. The required bias is approximately -6 volts. This bias should be variable to permit adjustment for changing tubes or other changes which would affect the plate voltage.

The same 300-volt power supply may be used on this stage as was used on the input stage. Since the plate is at 60 volts, there must be a 240 volt drop in the load resistor. Assuming a screen current of

30 μa in the following stage, the total current in this load resistor is 0.88 ma. The required resistance is then 273,000 ohms. In the amplifier which was constructed a 250,000 ohm resistor was used. The currents were therefore slightly higher than the values given above.

Following Stages

The third stage was built using 60 volts screen voltage, 90 volts plate voltage, and approximately -7.5 volts control grid bias. The supply voltage was the same 300 volts and the load resistor was 200,000 ohms.

When an attempt was made to put a fourth stage on the 300 volt power supply, undesired coupling through the power supply resulted in instability. It was therefore necessary to provide a separate power supply for all stages past the third. In order to permit the continued use of high load resistances and stepped plate voltages, this second supply voltage was made 400 volts. This supplied the fourth and fifth stages. The sixth stage required a third power supply for stability. The third power supply delivered 500 volts. It supplied the sixth stage and a 6V6 loading stage.

The operating voltages and load resistance values for all of the 6V6 stages are given in Table II. They were arrived at by a process similar to that given for the second stage. Values given for control-grid bias voltages are approximate since all bias voltages were made adjustable.

TABLE II

VOLTAGE AND RESISTANCE VALUES FOR 6V6 STAGES

Stage No.	2	3	4	5	6
E_{sg} volts	50	60	90	175	275
E_p volts	60	90	175	275	425
E_{g1}* volts	-6	-7.5	-12	-33	-60
E_{c1} volts	-22.5	-22.5	-22.5	-100	-100
E_b volts	300	300	400	400	500
R megohms	.25	.20	.30	.20	.20

*All values for E_{g1} are approximate since an adjustment was provided in all stages.

Another 6V6 was connected to the output of the sixth stage as a typical load. This tube was supplied from the same 500-volt power supply that was used on the sixth stage. Its load resistor was 1000 ohms. This loading stage could be arranged to operate a relay, a recording meter, or any one of a number of other possible loads. It was convenient because the low plate load resistor greatly facilitated frequency and other measurements.

Negative Feedback

Negative feedback from the plate of the second stage to the cathode circuit of the input stage was used. This feedback was useful for reducing drift introduced in the second stage. Since the drift compensation used in the input stage reduced the overall drift to the point where drift in the second stage was noticeable, this use of negative feedback was advantageous. The magnitude of the feedback resistor was adjusted experimentally to give a reasonable compromise between drift reduction and gain reduction.

POWER SUPPLY CONSIDERATIONS

Plate Supply

In any d-c amplifier, it is highly desirable to use regulated power supplies, especially on the initial stages. Unless a bridge balance or similar circuit is used a regulated power supply is almost a necessity. If many stages are to be supplied from one power supply, this supply must have a very low internal impedance (i.e., be very well regulated). Otherwise feedback resulting from this power supply impedance is likely to cause instability. It is also necessary to use a very stable bias voltage source.

For the amplifier described here all three of the plate supplies were regulated. They were available in the laboratory where the work was done and were used without any alteration to improve their stability or otherwise adapt them for this particular use.

Control-Grid Bias Voltage

For control-grid bias supply on the second, third, and fourth stages, the 22.5-volt tap on an ordinary 45-volt B battery was used. For the fifth and sixth stages, a higher bias was required than was available from a single battery. A regulated supply delivering -100 volts was available and was used for these latter stages. Two B batteries in series would have been satisfactory.

It would have been impractical to use an electronic supply for bias voltage on the first few stages because of the high degree of stability required. Since the control grid has a much higher gain than does the screen grid, any fluctuation of bias voltage would be much more serious than a similar fluctuation of plate or screen voltage. Battery supply was found to be satisfactory.

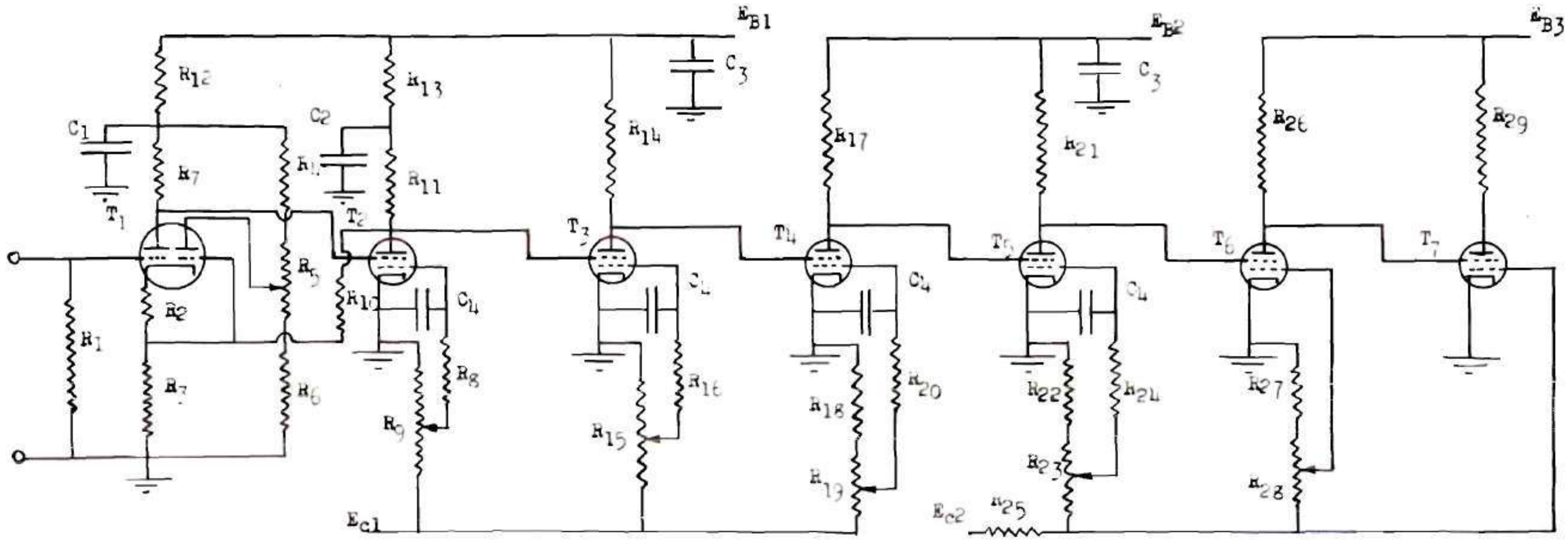
Heater Supply

A regulated 60 cycle supply was used for the heaters. This supply was sufficiently stable for use on all stages. However, in order to make some of the measurements required, it was necessary to use a d-c supply for the first three stages. This was necessary because of the large 60-cycle output which resulted from pick-up from the heaters and from other sources. This use of d-c for heater supply reduced the 60-cycle output to the point where the necessary measurements could be made.

The amplifier was not as stable with the d-c heater supply as with the regulated a-c supply. This was due to the fact that no d-c supply of sufficient capacity was available. Batteries were used when a d-c supply was necessary but after a short period the voltage would be reduced too much to be of further use. It was therefore necessary to use the a-c supply for the warm up period. After a sufficient warm up period the heaters could then be changed to the d-c supply for the purpose of making measurements.

BIAS VOLTAGES

It was necessary to make all control-grid bias voltages adjustable. This was done by using a separate bias supply bleeder with a potentiometer for each stage. These bias circuits can be seen on the complete circuit of Fig. 16.



Wire Wound Resistors

R_1	300,000 ohms
R_2	50
R_3, R_6	500
R_4	25,000
R_7	300,000
R_{10}	450,000
R_{11}	250,000
$R_{12, 13}$	1,000
R_{14}	200,000
R_{18}, R_{22}	6,000
R_{25}	3,300
R_{27}	10,000

Wire Wound Potentiometers

R_5	2,000 ohms
R_9, R_{15}, R_{19}	5,000
R_{27}	20,000
R_{23}	4,000

Carbon Resistors

R_8, R_{16}, R_{20}, R_{24}	100,000 ohms
R_{17}	300,000
R_{21}, R_{26}	200,000
R_{29}	

C_1	10	μf	C_3	1	μf
C_2	20	μf	C_4	.05	μf

T_1	6J5
T_2, T_3, T_4, T_5, T_6	6V6

E_{B1}	300 volts	E_{c1}	-22.5 volts
E_{B2}	400	E_{c2}	-100
E_{B3}	500		

Fig. 16. Complete Amplifier Circuit

In order to reduce the possibility of the circuit oscillating due to coupling through the control grids, each bias voltage was supplied through a low-pass RC filter. This filter does not affect a direct voltage but would reduce any higher frequency voltage in the control grid circuit.

PERFORMANCE

The gain of the amplifier was measured to be approximately 100 db. This gain is uniform from zero frequency (d-c) to over 1000 cycles and is down 3 db at 10,000 cycles. With the gain this high the drift is such that some adjustment must be made immediately before accurate measurements are to be made.

The amplifier was found to be very microphonic. To reduce this effect, the first tube (6J6) was mounted on sponge rubber and all connections to it were made with small flexible wire.

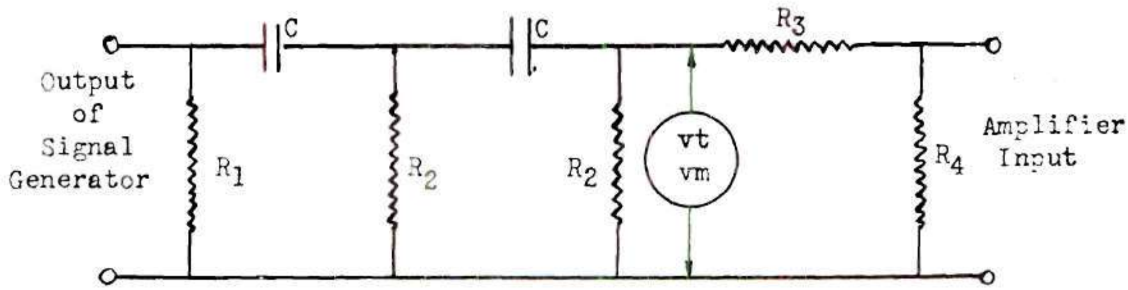
CHAPTER V

TEST DATA AND CONCLUSIONS

I. GAIN MEASUREMENTS

PROCEDURE FOR MAKING MEASUREMENTS

Gain measurements were made with a 1000-cycle signal from a Hewlett-Packard Model 200C Audio Oscillator. This oscillator was coupled to the amplifier input through a two-section high-pass R-C filter and a voltage divider. This circuit is shown in Fig. 17. The R-C filter served to reduce the 60-cycle component of the oscillator output signal. The voltage divider, consisting of two resistors, R_3 and R_4, reduced the signal voltage at the output of the filter by a factor of 100. This made



R_1	1,500 ohms
R_2	100,000 ohms
R_3	4,000 ohms
R_4	40.5 ohms
C	0.05 μf

Signal Generator - Hewlett-Packard Model 200C Audio Oscillator

Fig. 17. Signal Input Circuit

it possible to measure the voltage at the amplifier input with a voltmeter connected across the output of the filter. This was desirable because of the small values of signal voltage.

A Ballentine Model 300 Electronic Voltmeter was used to measure this input voltage. The deflection of this meter is proportional to the average value of a half wave of the A-C voltage. The scale is calibrated to give the root-mean-square, or effective, value of a sinusoidal voltage wave. Since the output of the oscillator was essentially sinsusoidal the readings may be taken as effective.

On the amplifier output connected from plate to ground were a General Radio Type 726A Vacuum Tube Voltmeter and a DuMont Type 208 Oscilloscope in parallel. The General Radio Voltmeter responds to the peak value of the impressed voltage. The scale was calibrated to give the effective value of a sinusoidal wave or 0.707 times the peak value of a complex wave. Because of the large 60-cycle voltage present in the output this meter was of little value in measuring the 1000 cycle signal voltage.

The procedure used for measuring the 1000-cycle output voltage was as follows. The oscilloscope screen was fitted with a transparent mask having ten grid lines to the inch. Then with a sinusoidal voltage with an effective value of 10 volts the oscilloscope amplifier was adjusted to give a deflection of one inch peak to peak. Thus the oscilloscope screen was calibrated in volts and could be used to determine the 1000-cycle component of voltage alone even though this 1000-cycle wave was superimposed on a 60-cycle wave.

GAIN

For the purpose of making gain measurements at 1000 cycles the output meters were connected to the plate of the final amplifier stage. This was more satisfactory than measuring at the plate of the load stage because of the larger voltages available. The load stage gave a voltage reduction of approximately ten to one.

Overall Gain

The amplifier was sensitive to voltages as low as 10 microvolts. The voltage gain was over 100 db for output voltages up to 20 volts. At output voltages much above 20 volts the output became distorted and above 30 volts clipping of the peaks occurred. The curve of output vs. input is shown in Fig. 18.

Gain of Individual Stages

Approximate values for the gain of the individual stages are given in Table III. This measurement was made by adjusting the output to 10 volts with the Ballentine meter on the plate of the sixth stage. Then the Ballentine meter was moved to the plate of the fifth stage and the output again adjusted for 10 volts. The ratio of voltages read on the Ballentine meter will then give the approximate gain of the sixth stage.

This method does not give accurate values because of the 60-cycle component of voltage. Since it is not known just where the 60-cycle voltage is introduced no correction can be made for it. It is likely that a small amount of this voltage is introduced in each stage. The measurement described will then give reasonably close approximations of the stage gains.

TABLE III

APPROXIMATE GAIN OF INDIVIDUAL STAGES

Stage No.	1	2	3	4	5	6
Gain	16	8	10	6	8	2

It may be noted that these stage gains are not as uniform as might have been expected. This is due principally to the tubes used. The performance of different tubes in this type of circuit may vary considerably. The tubes used in this amplifier were not picked to give maximum gain. A few tubes were rejected because of very poor current ratio, but in general the tubes used were average ones.

Overall Transconductance

For some applications the overall transconductance of an amplifier is of more interest than the voltage amplification. The transconductance of this amplifier from the input terminals to the output of the load stage can be easily determined. The net voltage amplification, including the load stage, is 10^4. The load resistance in the plate circuit of the load stage is 1000 ohms. Assume an input signal of 1 millivolt. The output voltage is then 10 volts. This 10 volts developed across a resistance of 1000 ohms requires that a signal current of 10 ma flow. Thus the transconductance, defined as output current divided by input voltage, is 10 ma divided by 1 mv, or 10 mhos. The load stage could be designed to give a higher transconductance if it is desired.

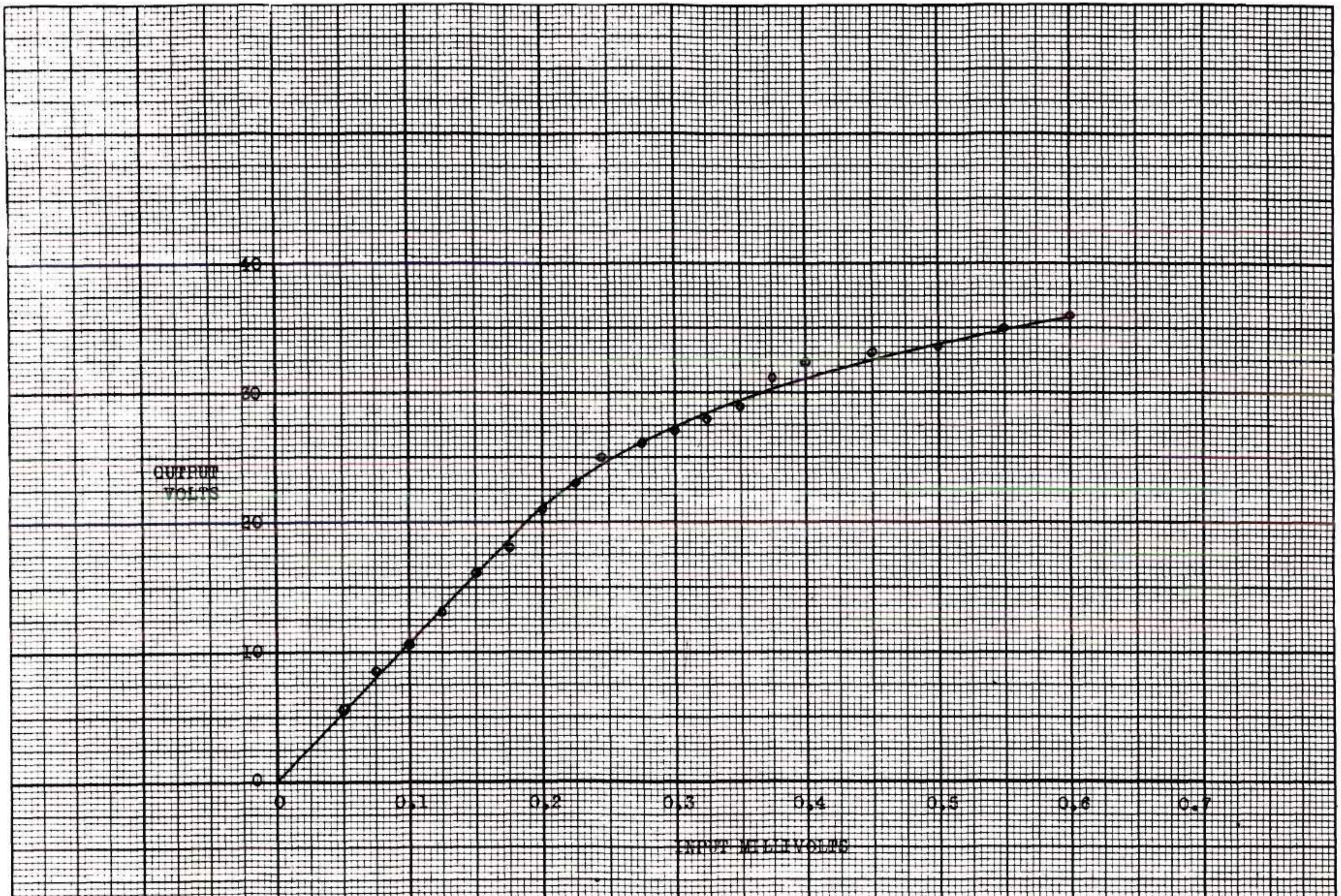


Fig. 18. Load Characteristic

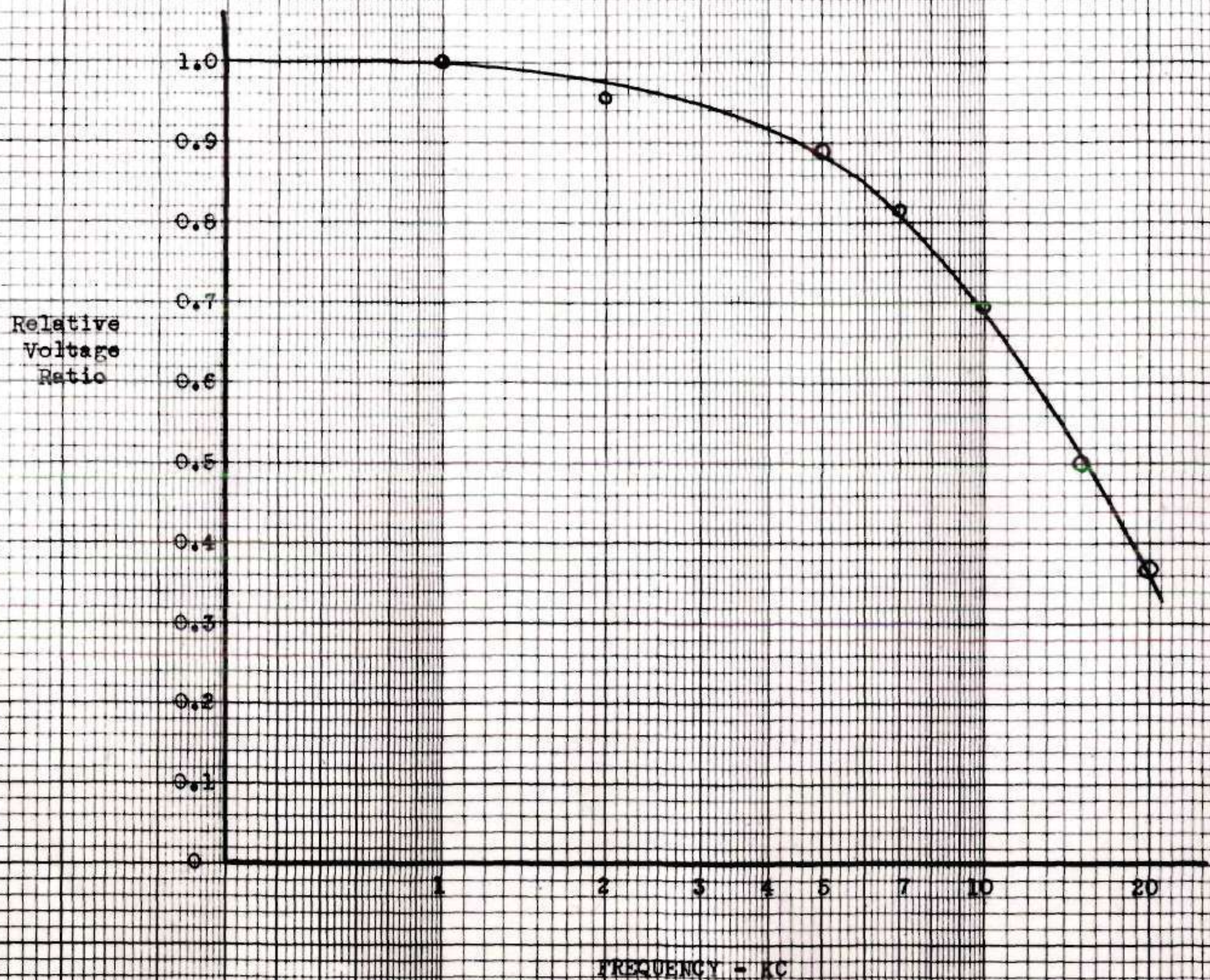


Fig. 19. Frequency Response

II. BANDWIDTH

The gain versus frequency characteristic is shown in Fig. 19. The amplifier gain is constant from zero frequency to over 1000 cycles and is down 3 db at 10,000 cycles. If slightly more bandwidth is desired it can be had by using smaller load resistors. This will of course reduce the gain. It would not be practical to increase the bandwidth appreciably.

III. DRIFT

With 100 db gain the drift is too great to permit use of the amplifier without some zero adjustment. Some of the drift observed was due to disturbances of the line voltage caused by machine shop equipment in a room adjacent to the laboratory. However, even under more ideal conditions, some drift would remain.

WARM-UP TIME

A period of several hours was required for the amplifier to warm up to the point where a steady drift due to the effects of temperature changes was not troublesome. During this warm-up period it was necessary to make adjustments of the bias voltages to keep the normal operating voltages at the plates and screens. If this was not done the amplifier would reach some equilibrium condition, usually with one or more tubes cut off. Then when the bias was adjusted to give the proper plate voltages the drift would again appear. The usual time required for warming up was from three to four hours during which time the constant attention of the operator was required.

ZERO ADJUSTMENTS

To permit adjustment of plate voltages to correct for drift, variable control-grid bias was provided for each 6V6 stage. During the warm-up period it was necessary to make bias adjustments on all stages to keep the plate voltages at their proper values. A Hickock Model 110A Vacuum Tube Voltmeter was used to observe plate voltages both during the warm-up period and during the operation of the amplifier.

After the amplifier had warmed up properly, correction for drift could be accomplished by making adjustments on only one stage. This was usually done on the fourth stage. The d-c voltmeter was connected to the plate of this stage and its voltage kept at the proper value by making small bias adjustments. An occasional check of the other plate voltages was made and adjustments made when they were required.

This adjustment could be made without the use of a d-c voltmeter. Then bias adjustments, preferably on only one stage, would be made to give maximum output. This method is not desirable since, after a short time, the d-c voltages may drift to a point where adjustment of one stage for maximum gain may not give the maximum gain available from the amplifier. Adjustment of several stages without a d-c voltmeter to indicate plate voltage is very unsatisfactory.

IV. SUMMARY

The purpose of this work has been the investigation of the possibilities of screen-grid coupling in d-c amplifiers and the development of an amplifier employing this type of coupling. This has been accomplished. The amplifier which was constructed had gain of over 100 db. Of this, 24 db were in the triode input stage and the remainder in the screen-grid coupled 6V6 stages.

Since the amplifier was constructed to demonstrate the use of screen-grid coupling rather than for any specific use, several refinements which might seem desirable were not considered necessary for the accomplishment of the original objective. For instance, three power supplies were used for plate supply, two for bias supply, and two for heater supply. It would, of course, be possible to build an amplifier with a less bulky assortment of power supplies. Also, there were approximately 8 to 10 volts of line frequency in the output. This was reduced to the point where satisfactory measurements could be made but it was not considered necessary to reduce it further.

The principal advantage of the screen-grid coupled amplifier is its simplicity. A common plate supply can be used for several stages. All cathodes are grounded. The only circuit element required for interstage coupling is the load resistor.

Without some further provision for reduction of drift, gain as high as 100 db is not considered practical for most applications. In general, applications requiring this much gain require more stability than was found in this amplifier.

It is probably best suited for applications such as automatic control circuits requiring gain of about 60 db. At this level the amplifier was reasonably stable after a much shorter warm-up time than was required with 100 db.

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APPENDIX I

GAIN OF THE CATHODE COMPENSATED STAGE

If the condition for compensation for cathode drift is satisfied, equation (12) of Chapter II becomes

$$\frac{i_1}{g_{m1}} = e_s - \frac{i_1}{1 + g_{m2}R_1}(R_1 + R_2) \quad .$$

Making the substitution $g_{m2} = 1/R_2$ this can be simplified as follows:

$$i_1 = g_{m1}\, e_s - \frac{i_1 g_{m1}}{1 + \frac{R_1}{R_2}}(R_1 + R_2)$$

$$= g_{m1}e_s - \frac{i_1 g_{m1}}{\frac{R_1 + R_2}{R_2}}(R_1 + R_2)$$

$$= g_{m1}e_s - i_1 g_{m1}R_2 \quad .$$

Solving this for i_1 we have

$$i_1 = \frac{g_{m1}e_s}{1 + g_{m1}R_2} \quad .$$

The output voltage is developed by this current flowing through the load resistance R_3. Thus we may write

$$e_o = i_1R_3 = \frac{g_{m1}e_sR_3}{1 + g_{m1}R_2} \quad .$$

The gain of the stage is then

$$\text{gain} = \frac{e_o}{e_s} = \frac{R_3}{\frac{1}{g_{m1}} + R_2} \ .$$

As was noted when g_{m1} was first introduced (equation (5) Chapter II), this is the g_m as measured along the dynamic load line, not the more commonly used g_m in which the plate voltage is considered to be constant. The change in plate current caused by a small change in grid voltage of T_1 is

$$\Delta i_{p1} = \frac{\mu \Delta e_{g1}}{r_{p1} + R_3 + R_2} \ .$$

The operating g_m is then given by

$$g_{m1} = \frac{\Delta i_{p1}}{\Delta e_{g1}} = \frac{\mu}{r_{p1} + R_3 + R_2} \ .$$

Making this substitution into the expression for gain we have

$$\text{gain} = \frac{R_3}{\frac{r_{p1} + R_2 + R_3}{\mu} + R_2} \ .$$

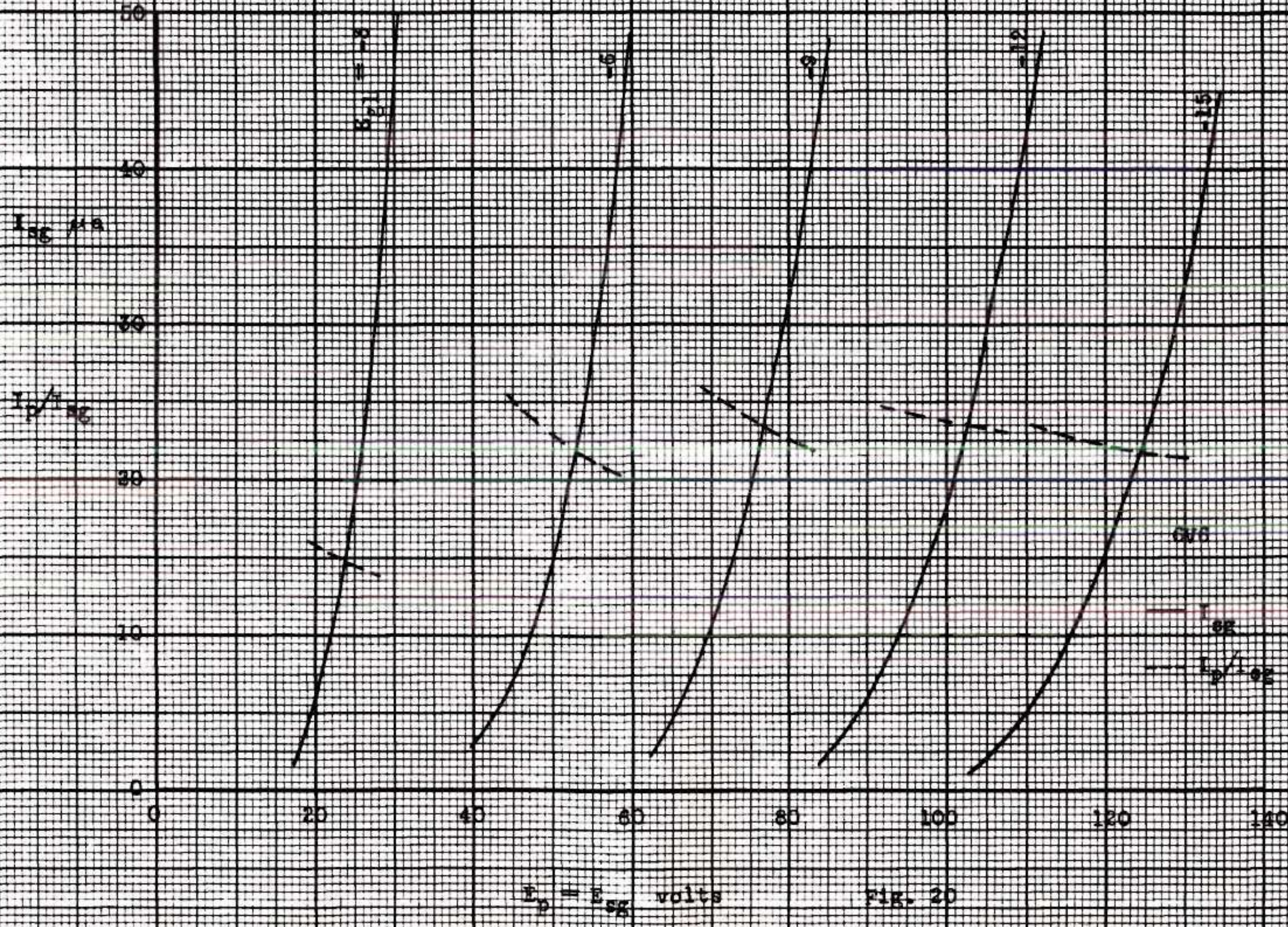
This may be reduced to

$$\text{gain} = \frac{\mu R_3}{r_{p1} + R_3 + R_2(1 + \mu)} \ .$$

The cathode resistor R_1 does not enter into the gain equation because it does not introduce negative feedback to T_1. The compensating tube T_2 compensates for any voltage developed across R_1 just as it compensates for the cathode drift voltage v. The only effect R_1 has on T_1 is that it furnishes a constant bias voltage.

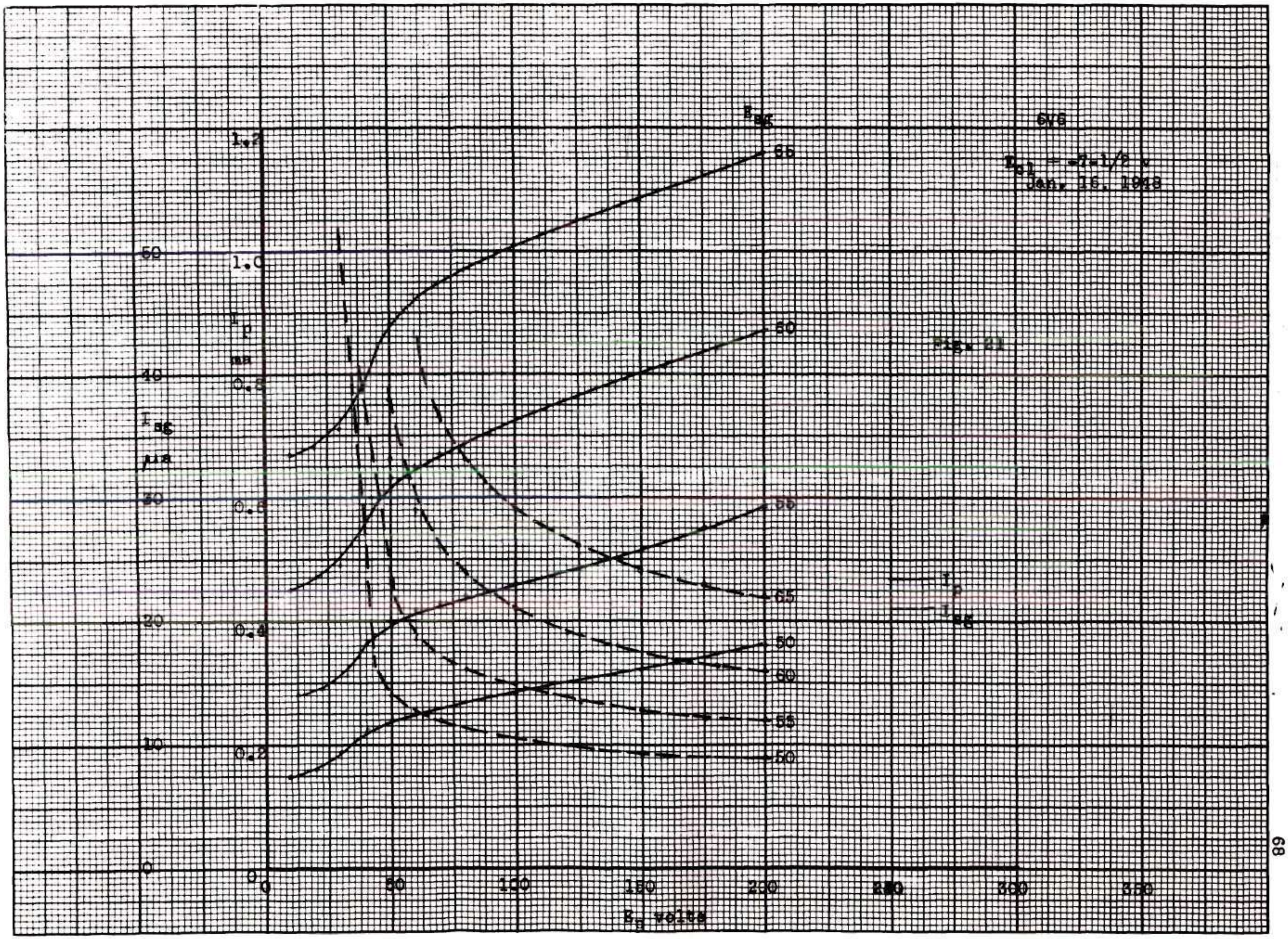
APPENDIX II

6V6 CURVES AND DATA



I_{sg} μa
I_p/I_{sg}
0
10
20
30
40
50
0
20
40
60
80
100
120
140
$E_{g1} = -3$
-6
-9
-12
-15
6V6
— I_{sg}
--- I_p/I_{sg}
$E_p = E_{sg}$ volts

Fig. 20


6V6
E_g1 = -7-1/2 v
Jan. 16, 1948
Fig. 21
E_sg
65
60
55
50
I_p
ma
I_sg
μa
I_p
I_sg
1.2
1.0
0.8
0.6
0.4
0.2
0
50
40
30
20
10
0
0
50
100
150
200
250
300
350
E_p volts

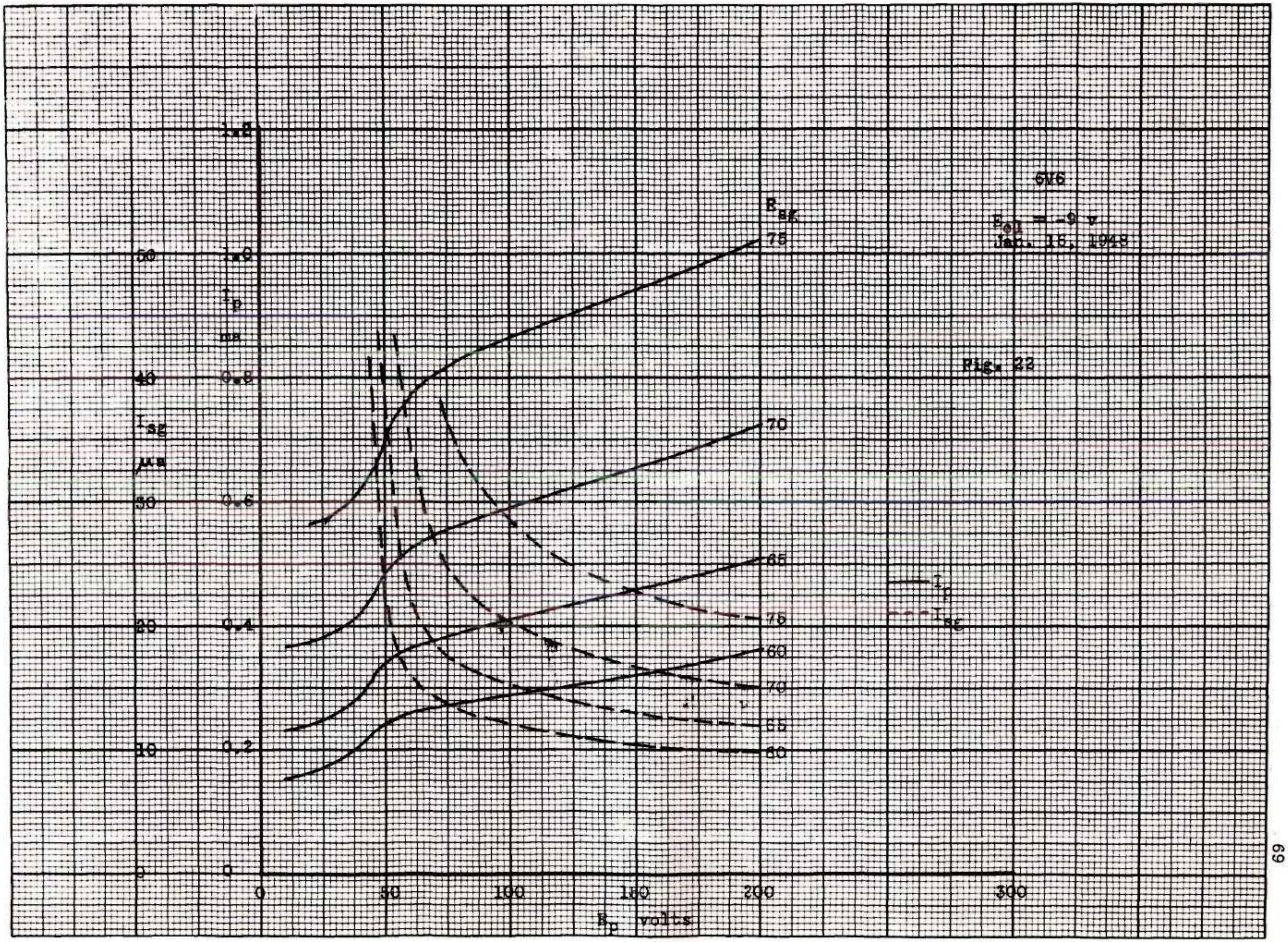

6V6
E c1 = -9 v
Jan. 16, 1948
R sg
75
70
65
60
75
70
65
60
1.2
1.0
0.8
0.6
0.4
0.2
0
I p
ma
50
40
30
20
10
0
I sg
μa
0
50
100
150
200
300
E p volts
I p
I sg

Fig. 22

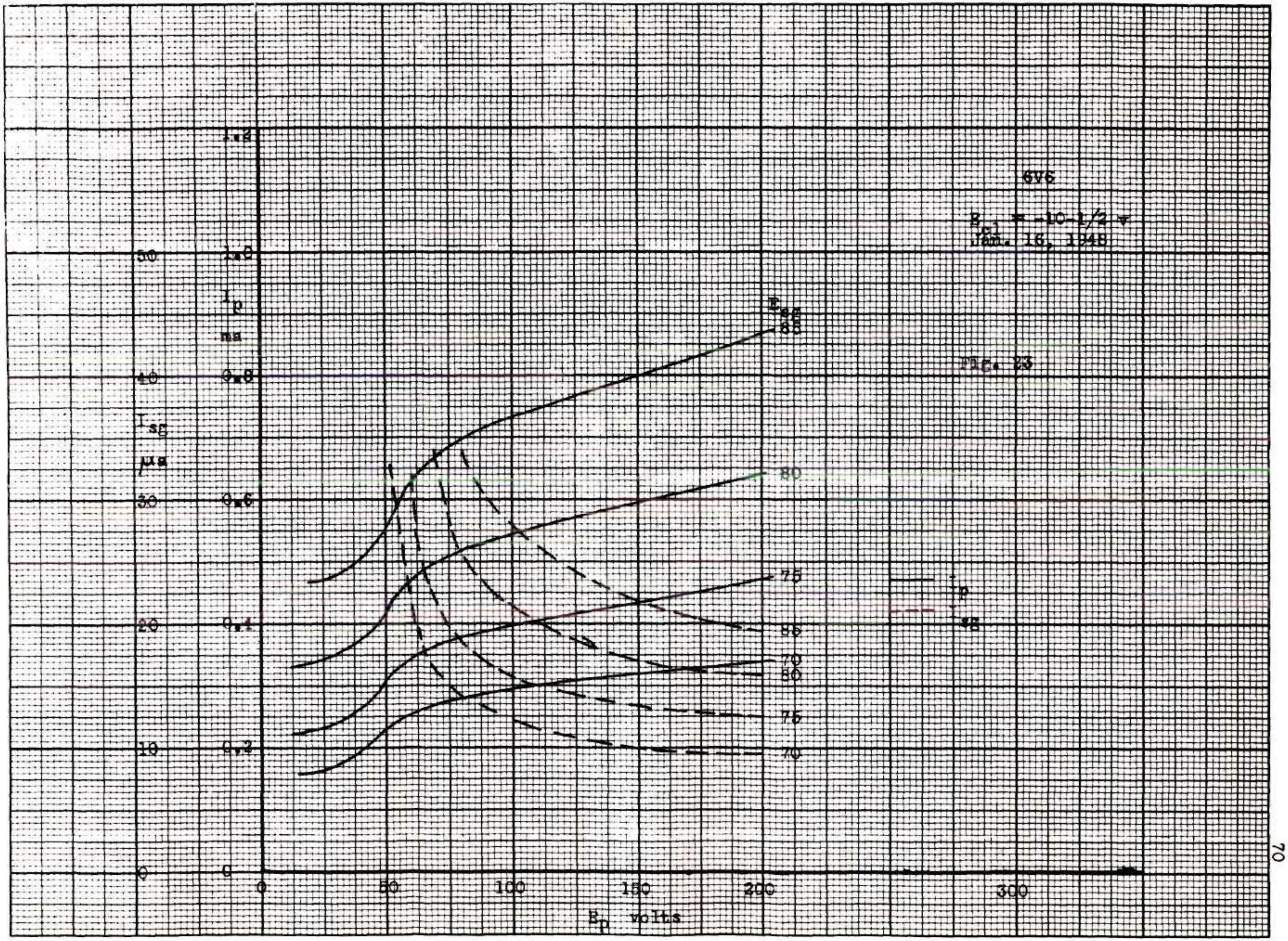

6V6
Ec1 = -10-1/2 v
Jan. 16, 1948
Fig. 23
Ip
ma
Isg
μa
Esg
85
80
75
70
85
80
75
70
Ip
Isg
1.2
1.0
0.8
0.6
0.4
0.2
0
50
40
30
20
10
0
0
50
100
150
200
300
Ep volts

TABLE IV

DATA ON A TYPICAL 6V6

E_{sg} volts	E_p volts	I_p ma	I_{sg} μa	E_{g1} -volts
50	50	.84	44	6.1
	60	.85	30.5	6.2
	70	.84	25.5	6.2
	80	.86	24	6.3
60	70	.86	31.5	7.4
	80	.87	28	7.4
	90	.86	25	7.5
70	80	.86	33	8.7
	90	.87	29.5	8.7
	100	.89	27.5	8.7
	100	.86	26	8.8
80	90	.85	31	10.2
	100	.86	29.5	10.2
	110	.88	27	10.2
	120	.90	25.5	10.2
	120	.85	24	10.3
90	100	.86	32.5	11.8
	110	.88	30	11.8
	120	.89	28	11.8
	130	.91	26	11.8
100	110	.86	32.5	13.3
	120	.88	31	13.3
	130	.90	29.5	13.3
	140	.91	27	13.3
	140	.86	26.5	13.5
110	120	.86	32.5	15.0
	130	.87	31	15.0
	140	.88	29	15.0
	150	.89	27	15.0
	150	.86	27	15.1
120	130	.87	33	15.8
	140	.88	31.5	15.8
	150	.89	30	15.8
	160	.90	29	15.8
	160	.86	27.5	15.9
130	140	.85	32	22.5
	150	.86	31	22.5
	160	.87	30	22.5
	170	.88	29	22.5
140	150	.87	34	25.0
	160	.88	33	25.0
	180	.90	31	25.0

TABLE IV (cont.)

DATA ON A TYPICAL 6V6

E_{sg} volts	E_p volts	I_p ma	I_{sg} μa	E_{g1} -volts
150	160	.86	34	27.0
	170	.87	32.5	27.0
	180	.88	32	27.0
	190	.89	31	27.0
160	170	.85	34	29.6
	180	.86	33	29.6
	190	.865	32	29.6
	200	.87	31	29.6
170	180	.86	35	32.0
	190	.87	34	32.0
	200	.875	33	32.0
	210	.88	32	32.0
180	190	.86	35	34.5
	200	.87	34	34.5
	210	.88	33.5	34.5
	220	.885	33	34.5
	230	.89	32.3	34.5
190	200	.86	35.5	37.2
	210	.865	35	37.2
	220	.87	34	37.2
	230	.875	33.5	37.2
	240	.88	33	37.2
200	210	.86	36.5	40.0
	220	.865	35.5	40.0
	230	.87	34.8	40.0
	240	.875	34.2	40.0
	250	.88	33.6	40.0